ANALYSIS OF THE PARAMETERS REQUIRED FOR PERFORMANCE MONITORING AND ASSESSMENT OF MILITARY COMMUNICATIONS SYSTEMS BY MILITARY TECHNICAL CONTROLLERS

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THESIS

Analysis of the Parameters Required for
Performance Monitoring and Assessment
of Military Communications Systems
by Military Technical Controllers

by

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Thesis Advisor: Eugene J. Normand

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20. Abstract

from which these parameters are derived from are: level, noise and distortion.

The unsuitability of current instruments to provide quality assurance of the numerous circuits appearing at technical control facilities is discussed and a dedicated, task oriented, combination measuring set is evaluated as to its ability to measure the parameters that have been determined.

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I. INTRODUCTION

The military communication complex as it exists today is a heterogeneous mixture of equipment assembled from the military and commercial inventory to perform many interrelated functions. The individual equipment has been installed over an extended period with a minimum of system engineering. The present system is a loose aggregrate of equipment tied into the total complex by whatever matching devices could be acquired. A typical long haul communications system might be as illustrated in Figure 1. In order to achieve some flexibility and communications capability a manual "brute force" method was devised to interconnect each of the individual transmission media in a facility called "technical control."

The rapid growth and increasing sophistication of military communications has now pointed to a need for more effective technical control. The implementation of highly sophisticated systems such as Autodin, Autovon, Autosevocom, and the deployment of high density transmission systems require a change in the management and control philosophy.

One area that is being considered by all services is the effort to develop a more automated technical control. Many studies have been conducted into the feasibility of automated technical control. Many more studies are being conducted at the present time. The emphasis in all of these studies is on hardware alone, seldom is there any reference to upgrading the level of competence of the technical control personnel. Automation is being designed with the intention of increasing the operational effectiveness of the TCF, but this cannot be done by improved hardware alone.

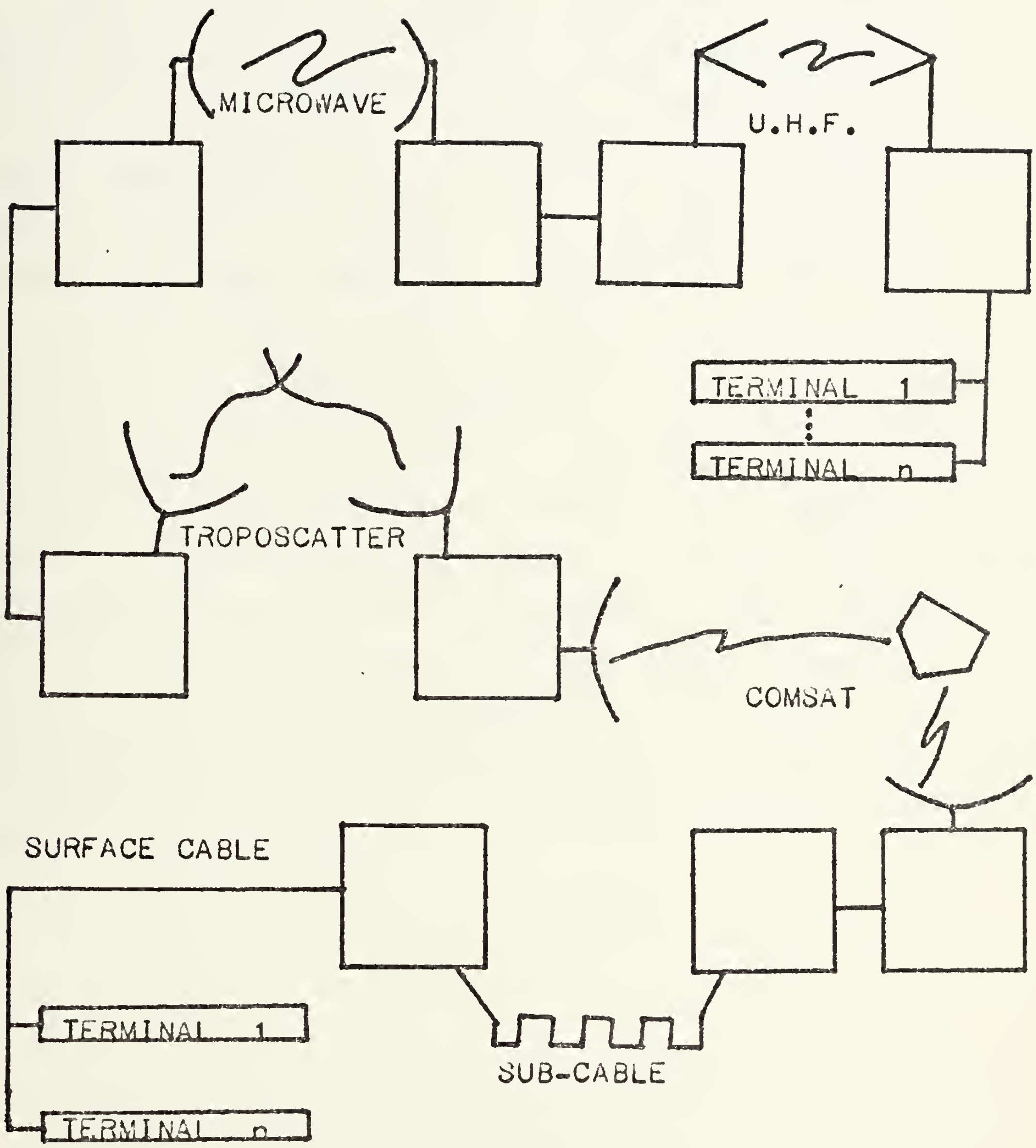


Figure 1- Typical Military Communications System

(Long-Haul, Wide-Band, Multi-Channel)

There appears to be a distinct lack of training, skill, and knowledge concerning modern communication equipment and systems among technical control personnel.

One problem that exists is the determination and explanation of the performance monitoring and assessment parameters that technical control personnel will be required to know and understand. It is the purpose of this thesis to analyze the military communication system to determine and explain these performance monitoring and assessment parameters. After sufficient background on the basic nature of military communications systems, military technical control, and the key communication system performance indicators, each of the parameters will be discussed in significant depth so as to determine its importance in relation to performance monitoring and assessment. It should be understood that while an in-depth study could be conducted on each of these parameters, that such is not possible within the length of this thesis. Again, the goal is to develop a set of parameters that if measured and analyzed correctly will allow military technical controllers to adequately monitor and assess the performance of military communications systems.

II. MILITARY TECHNICAL CONTROL - PRESENT PROBLEMS AND FUTURE TRENDS

To provide maximum communications capability, a method is required to anticipate difficulities and permit appropriate corrective action before actual failure of the communications system. This function is assigned to technical control under the classification of a quality assurance program. This function is attempted by measuring certain parameters, and based upon analysis of these parameters it is assumed that a prediction of circuit failure may be made before total failure occurs. This method requires; selection of the correct parameters to examine, proper and timely measurements of these parameters, and intelligent analysis of these measurements to predict failure.

The existing philosophy and concept of technical control is defined in DCAC 310-70-1 (References 1-7). The basic defination and functions of technical control are given in Appendix A. It is interesting to note that the original illustrations in the 1955 publication are still being published, which clearly indicates the lack of overall progress. The present DC control consoles, with their associated hit counters, can be misadjusted and misused and are of only limited value today. The methods for analog or voice channel analysis are most basic. There are no methods proposed or stated requirements to provide on-line checks of the complex composite waveforms being generated by the latest modems.

The unsuitability of the current methods to provide quality assurance of the numerous circuits appearing at large technical control facilities has been operationally

demonstrated. This statement is supported by recent studies which clearly indicate that the measurements themselves are not a valid basis for judging circuit quality. The current techniques of go or no-go circuit philosophy can never permit assessment of impending failure, but only prove that a circuit has failed. These procedures result in unsatisfactory customer service and adversely impair the overall communications mission.

Since technical control methods and procedures have not kept pace with recent developments in communications technology, studies have been undertaken to upgrade technical control functions, including the feasibility of utilizing automated techniques and procedures. Experience has shown that even limited automated assistance resulted in significant improvements of technical control and system performance. Some studies along this line (References 8-16) are:

The Air Force Concept for Semi-Automated Technical Control (SATEC)

Automatic Facilities Report

Army Automated Quality Monitoring Reporting System (AQMRS)

Army Automated Technical Control-Semi (ATC-Semi)

DCA Technical Visits Program (TVP)

AFCS Scope Creek and Link Performance Assessment Programs (LPA)

The Navy Automatic Digital Data Analyzing System (ADDAS)

The Joint Tactical Communications Office (Tri-Tac)

Concept of Tactical Communications Control

The Post-1980 Facilities Control Concept Study by the

Army Electronics Command (ECOM)

Elements of Automated Technical Control (ATEC) now under development will provide assistance to the controller, either by independently operated monitoring tests and analysis of system components in individual communication terminals, or in conjuction with data processing computers. The processing units are essentially high speed digital computers with memories and other peripheral facilities adequate to perform the automated tasks in support of the controllers. Those automated systems under development are essentially designed to:

A. Monitor critical equipment operation alarms which may be indications of conditions leading to failures.

B. Perform in-service tests on a continuous basis while the system is in operation and without disrupting service.

C. Perform out-of-service tests, which necessitate taking the system or parts of it out of regular service for the duration of the test.

D. Recognition, organization, recording, analysis, and reporting of test results.

The primary function of technical control then becomes equipment status monitoring. All the major equipment in a system would have internal sensors with properly selected parameters monitored , so that operational performance of the system could be accurately assessed from these measurements. (see Figure 2)

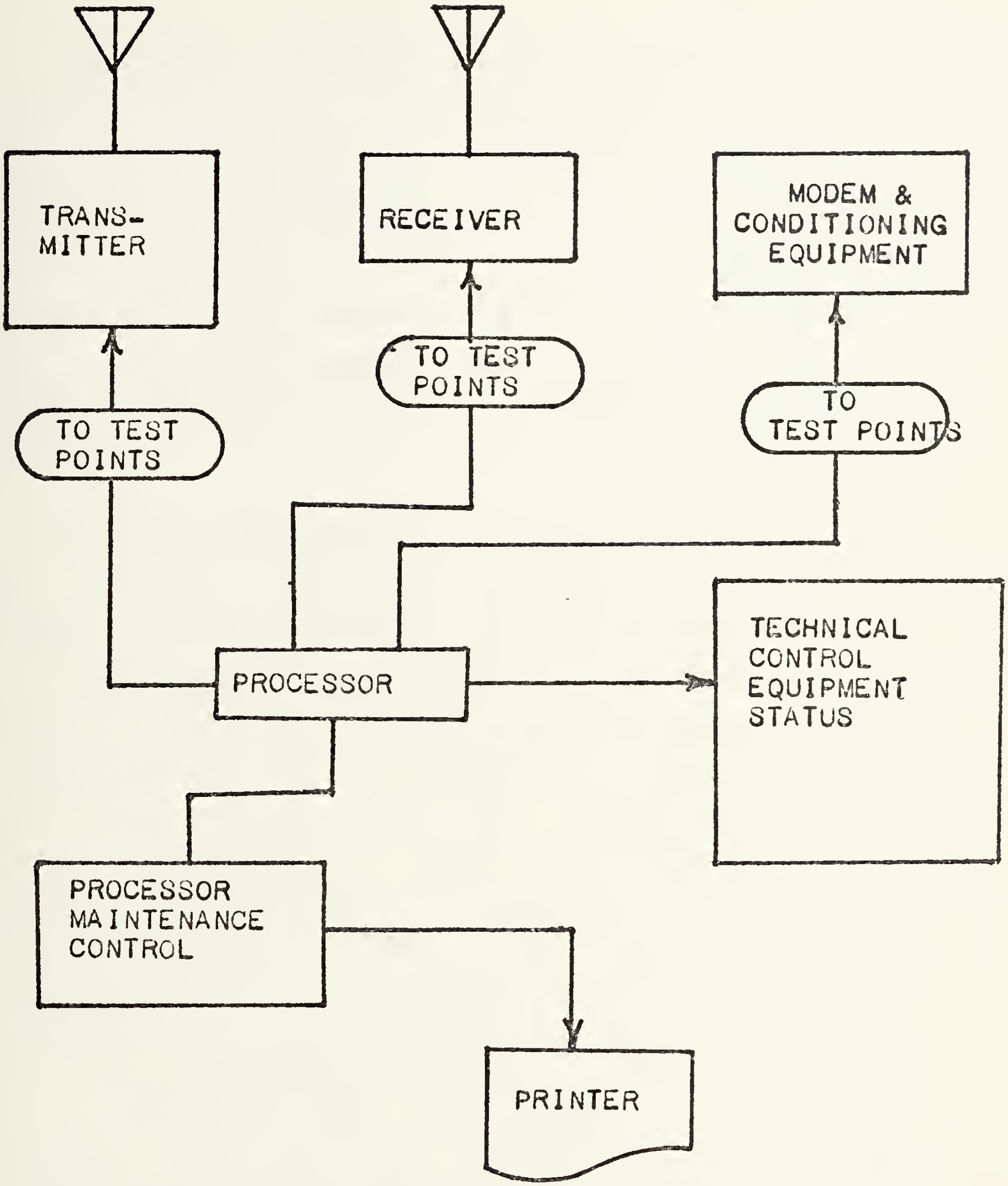


Figure 2- Equipment Status Monitor

Performance assessment would include using the system performance data to inform the technical controllers of the system's "health" and predicting impending system degradation or failures. The automated capability will allow the technical controller to check channel, link, or system performance in near real-time; he will be able to isolate faults in seconds. The technical controller will then be able to maintain optimum system performance, and will be able to initiate and manage meaningful system control.

As can be seen automated technical control will be a giant step in military communications. However, fully automated technical control facilities are costly and will not be practical at lower levels of communications. What is needed is a meaningful manual quality assurance program at an intermediate level (shipboard or division). There have been studies along this line (NAVCOMMCOM Instruction 2300.13, Communication Quality Monitoring and Control Program, and NAVTELCOMM Instruction C2300.19, Shipboard Quality Assurance Program, References 17 and 18) but they are not up-to-date or even very realistic. The tests and measurements are out of date and of little practical use in evaluating communications systems. There remains one basic problem, however, that the studies to date seem to have neglected. That is the determination of meaningful parameters required to measure and analyze the communication system. The remainder of this thesis will be devoted to this problem.

III. KEY COMMUNICATION SYSTEM PERFORMANCE INDICATORS

A. GENERAL

There are many different symptoms of trouble that may occur in communications circuits. Some of these symptoms are considered "key" indicators since they indicate the performance of the system as well as the performance of the individual circuits. These key indicators are levels, noise, and distortion. These indicators all interact with one another and with other performance parameters. Examination of these three indicators can offer an insight into the over-all performance of the system.

B. LEVEL

Level is an expression of relative signal strength at various points in a communications circuit. In communications terms, the level is given in dB relative to a reference level. Usually 1 milliwatt is the reference used and the unit becomes dBm. Test Level Point (TLP) is measured in dBm and establishes a reference level from which test levels may be reckoned. A level given in dBm0 is given in reference to the TLP.

$$\text{dBm0} = x\ (\text{dBm}) - \text{TLP}\ (\text{dBm})$$

Some point in the circuit is arbitrarily chosen as a reference point for signal level measurements. The relative level at any point in a circuit is a measure of the power gain or loss between the 0 TLP and the point under consideration. Signal powers and interference levels may be referred to the 0 TLP as "a signal level of -16 dBm0," which indicates the power level the signal would have registered had it been measured at the 0 TLP. This arrangement serves a number of purposes. It permits establishment of specified

losses along a circuit. With 0 dBm sent at the TLP, the circuit can be adjusted to obtain the proper levels at other TLP's, and the losses are thereby adjusted. For example, the inserted loss between a 0 TLP and a -4 TLP must be 4 dB. To produce a proper level at a +2 TLP, 2 dB of gain must be inserted. By the use of TLP, circuits can be maintained so that each will operate at the correct level. System transmission levels may be referred to TLP's. When a station transmits a signal measured at -8 dBm at the 0 TLP, the transmit level is -8 dBm0. Noise levels may also be established by reference to a TLP. A transmission system requiring a 30 dB signal-to-noise ratio requires that noise at the 0 TLP not exceed -30 dBm.

For direct circuits, the transmitter is considered to be the 0 TLP. The TLP at the receiver is established by the overall loss of the circuit. When the circuit consists of a multiplex channel, other TLP's are involved. Multiplex channels are usually lined up to provide 23 dB of gain in each direction. The transmitting point is a -16 TLP, and the receiving point is a +7 TLP. Other parts of the circuit are arranged to accomodate these levels. All classes of switching centers are considered to be -2 TLP's. The point of test is theoretically at the center of the switch.

Level measurements are used to determine the loss and attenuation distortion of a circuit. In addition, level measurements of the received level of a signal are used in conjunction with noise measurements to determine the signal-to-noise ratio of the circuit. Some level measurements of importance are:

Loss which is the end-to-end circuit attenuation, usually measured at 1000 Hz.

Attenuation distortion is the loss deviation (1000 Hz reference) over the range of frequencies of interest. This consists of static and dynamic frequency response and bandwidth.

Return loss is a measure of the mismatch between the actual communication circuit impedance as compared to a nominal defined impedance.

Long term loss variations are changes in the loss of a circuit due to aging of components, changes in physical makeup, and temperature variations.

Baseband loading is the result of high levels in the baseband of multiplexed circuits. The radio and multiplex equipment are usually designed to operate near the system overload point at peak traffic periods. This insures optimum over-all performance. However, this assumes that each user's input is at a certain level and that the system is aligned properly. Any high levels in the baseband raise the loading level to the modulator and can cause severe intermodulation noise in all channels. (see figure 3)

Multiplexers are designed for certain levels of loading, if these levels are exceeded then intermodulation noise can develop. There is a real tendency among subscribers to believe that if they raise the level, the signal will get there louder and better. This is not valid if raising the level generates more noise and distorts the signal.

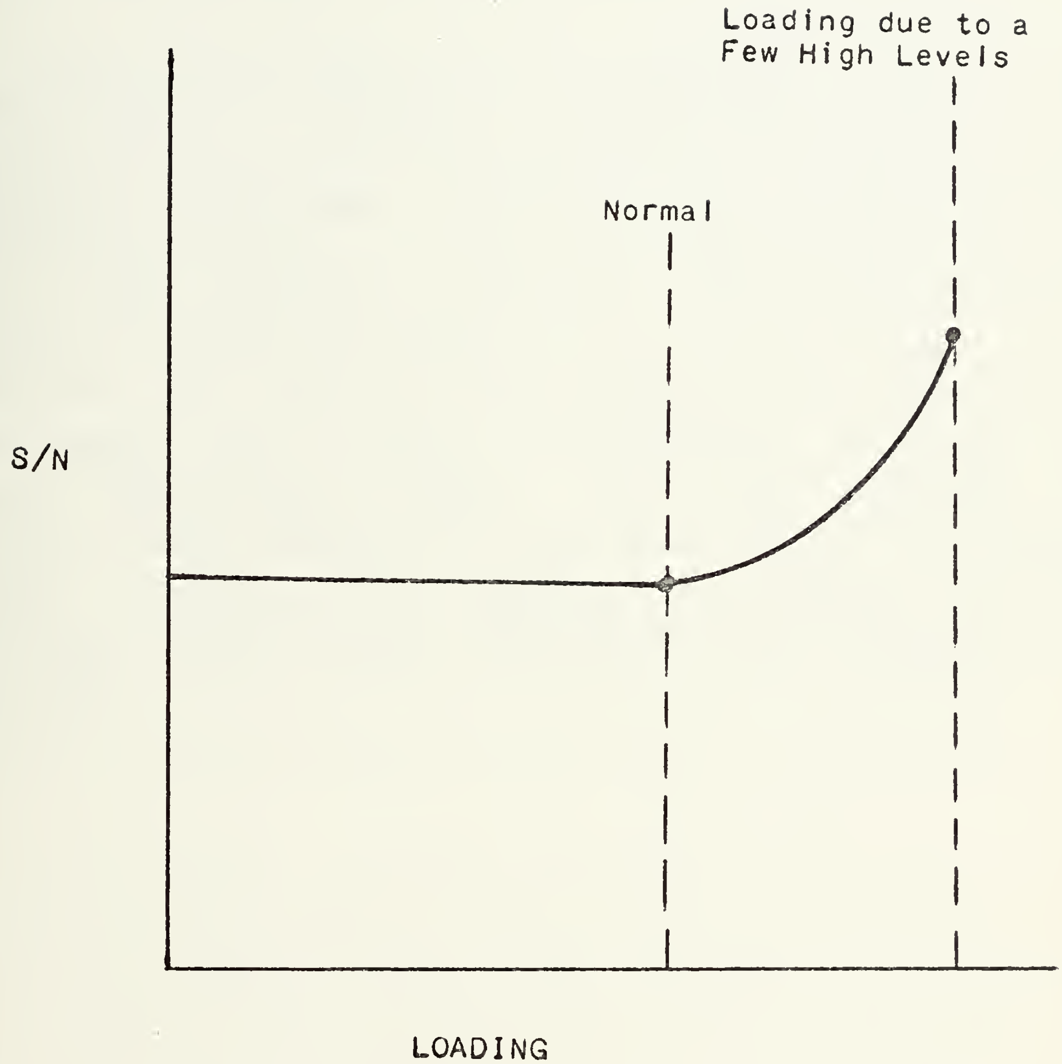


Figure 3- Effects of a Few High Levels on Channel S/N

Baseband loading is referenced to the test point level. The actual value of loading is measured in dBm absolute with a meter registering a true RMS reading.

Level measurements may be performed in-service or out-of-service. In-service measurements involve the measurement of the level of the traffic on a circuit without disrupting subscriber service. This offers a good performance indicator, in that if the level of the signal is not at the standard transmission level for the type of traffic, a problem is suspected to be present. There is no substitute for effective level discipline. It is essential that correct signal levels be maintained throughout the entire communication system. It is not sufficient to correct a level problem at a point other than the problem source. Levels can be measured and adjusted for each individual channel, group, supergroup and transmitter or receiver.

C. NOISE

Any signal which carries information has one essential characteristic; that of an ordered structure. To establish communications, it is necessary to maintain that ordered structure over the signal path. Since randomness is a more natural state than order, the signal is always under disruptive influences. The major impairment of electronic communications signals is noise. As used here, noise refers to all signals which are random in nature and mask the actual communications signal.

Thermal noise is inherent in nature. All objects radiate energy over a wide range of frequencies. This radiation, in turn, appears as noise to a radio antenna. Thermal noise is also generated in electrical circuits by electrons randomly colliding with each other during current flow through a conductor. These processes add to produce a composite level of thermal noise often called "white" noise.

The amount of thermal noise power emitted by a source is directly proportional to the absolute temperature of the source and the bandwidth of the noise under consideration. For example, a system with 10 MHz bandwidth admits twice as much thermal noise as a system with a 5 MHz bandwidth. Similarly, an antenna pointed at a satellite in space would see far less thermal noise than one pointed along the earth, due to the low radiating temperature of space. The two major classes of thermal noise commonly found in communications systems are "front-end" and "idle" noise.

The first class is generated in the front-end circuits of the receiver. Due to automatic gain control (AGC) circuitry in the receiver, front-end noise varies inversely

with RF signal strength. As the signal strength increases, the gain of the first stages is suppressed hence, noise generated in the front-end circuits is amplified less. Fading of the RF signal will cause an increase of noise in the system.

The second class, usually referred to as idle noise is generated by the transmitter circuitry and later portions of the receiver circuitry. Since this noise is added by stages whose gain is not variable, it is not affected by RF signal strength. Thus, it represents an absolute limit on the noise performance of the system.

The thermal noise contribution of any device can be expressed in terms of a noise figure. Noise figure is defined as the amount of noise added by a device as compared to an ideal device. The noise figure of a device depends on the temperature and bandwidth of the applied noise. Noise figure is usually defined assuming a temperature of 290 degrees K (63 degrees F) unless otherwise specified.

Impulse noise, in contrast to thermal noise, is sporadic in nature and occurs in bursts. It consists mainly of discrete, high amplitude pulses of short duration and is caused by lightning aurora, ignition noise, power lines and associated switching equipment, and dialing impulses in telephone systems. Speech signals are virtually immune to disruption by impulse noise. A speech sound is sustained over a period of time and noise impulses are too brief to have an effect on the intelligence present in the sound. While having little effect on speech, impulse noise presents a great problem for data. There is no redundancy in a data pulse as there is in speech. Impulses can easily be mistaken for data pulses. A short burst of impulse noise can turn a data stream into a meaningless jumble. In Figure 4, impulses crossing the decision level are counted as data

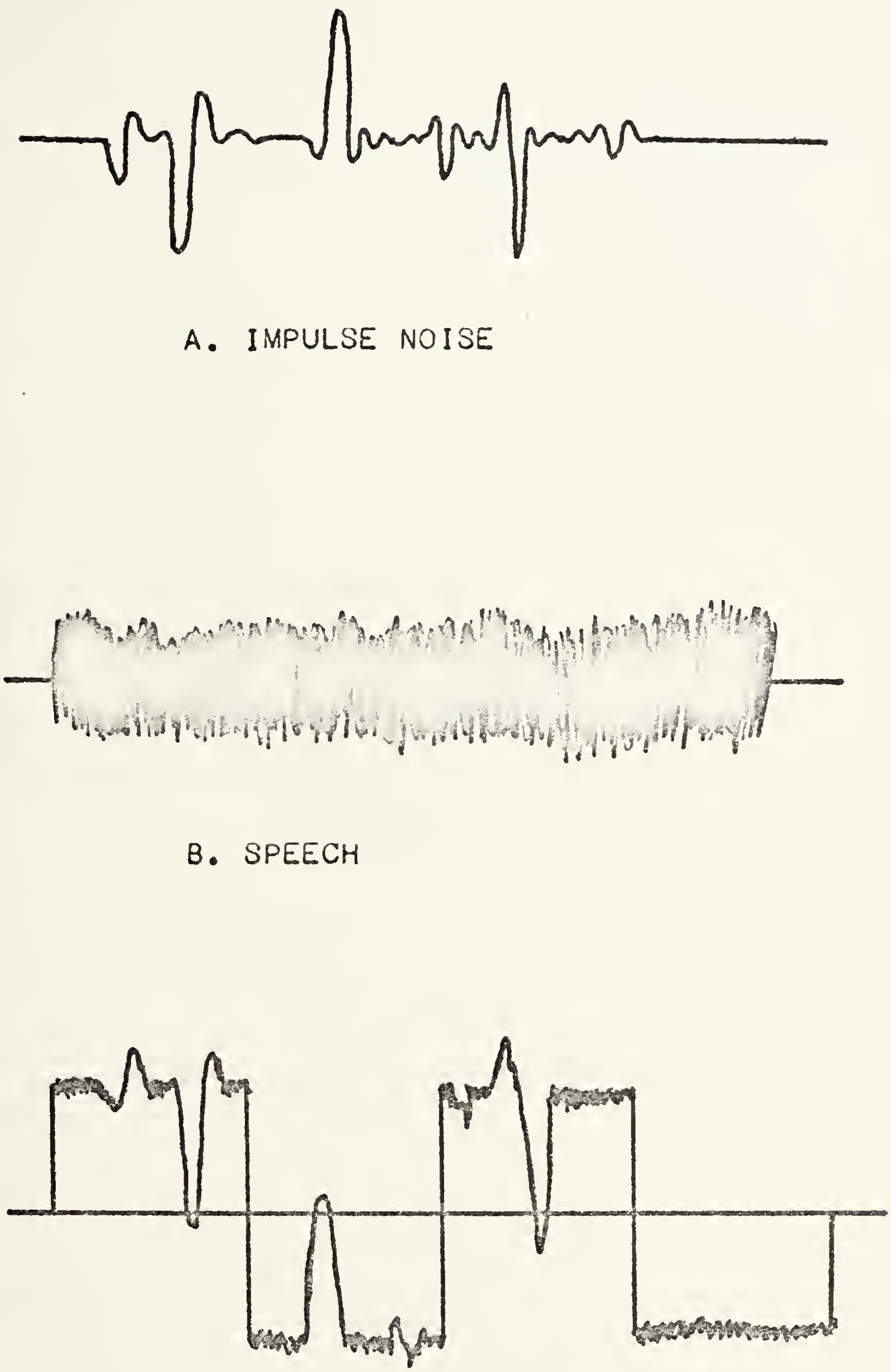
pulses. Note that white noise below the decision level does not affect data. Even if the impulses are much shorter in duration than the data pulses, the impulses can cause tuned circuits to ring and interfere with the data signal.

Crosstalk is the "leaking" of a signal from its allotted channel into other channels. In wideband systems, crosstalk is generated in the multiplex equipment. Adjacent channels in frequency division multiplex are separated by filters. If signal levels become excessive or if the filters are not selective enough, some signals from an adjacent channel may appear in another channel.

Crosstalk is not the only way a channel can interfere with other channels. Nonlinear operation of any element in the system will cause spurious signals to be generated. The effect of nonlinear operation on a single frequency can be seen in figure 5.

If more than one frequency is passed through the nonlinear element, harmonics of all the frequencies present, plus the sums and differences of all frequencies and their harmonics are present at the output. Second order intermodulation products consist of sums and differences of two frequencies (A+B, A+C, A-B,etc.). Third order products are combinations of three frequencies (A+B+C, A+B-C 2A+B, etc.). Higher order products appear farther away from the fundamental frequencies. Hence, intermodulation generated in a channel can interfere with channels far removed in frequency from the interfering channel.

In normal operation, the power in the intermodulation products increases with increasing input level. Initially,



A. IMPULSE NOISE

B. SPEECH

C. EFFECTS OF IMPULSES ON DATA

Figure 4- Impulse Noise, Speech, and Data Pulse

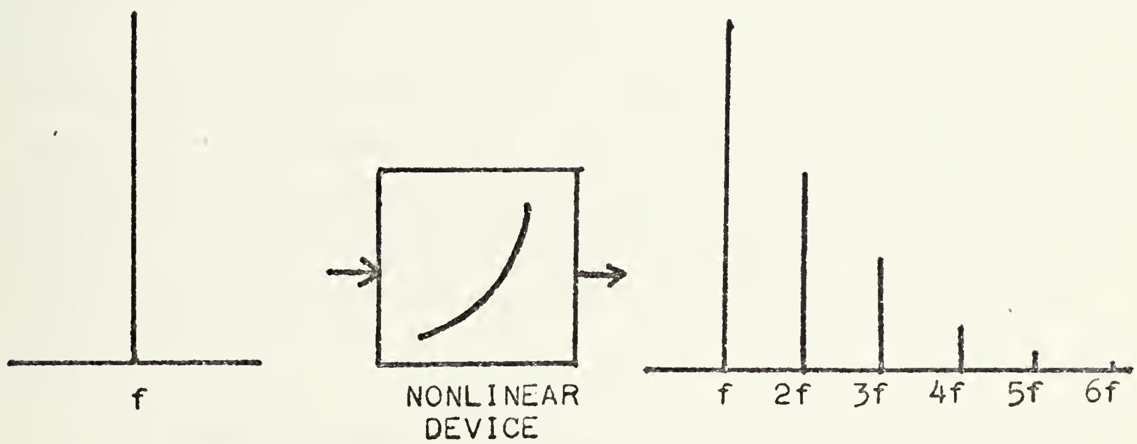


Figure 5.- Harmonics Due to Nonlinear Operation

second order products have the greatest amplitude, third order products the next greatest, and so forth. However, a 1 dB increase in input power causes a 2 dB increase in second order products, a 3 dB increase in third order products, etc. This situation continues until the system breakpoint, or overload point is reached. Beyond this point, the intermodulation power rises rapidly (as much as 20 dB for a 1 dB increase in input power). Higher order products rise more rapidly and eventually can equal the 2nd and 3rd order products in magnitude. Thus, sizeable disturbances may occur in nearly all channels in a system due to severe intermodulation in one channel. The intermodulation products of a complex signal such as speech are so widespread that they resemble noise in their randomness. Intermodulation noise on a wideband system appears as an increase in the general noise level. This topic will be discussed in greater detail in the section on distortion.

Two other noise parameters to consider are single frequency interference and quantizing noise. Single frequency interference is spurious tones present on the channel in addition to the desired signal. Quantizing noise is signal correlated noise generally associated with the quantizing error introduced by analog-digital and digital-analog transmission systems.

There are many different units needed to measure noise. This need has been caused by; the many types of noise present within a system, the types of handsets or reproducers that have been in common use (reproducing the noise as well as the speech), and the different ways in which noise affects modern day communications equipment (data, facsimile, voice, etc.). Noise in a voice channel can assume many forms, for example, white noise, impulse noise, discrete noise, etc. The disturbing effect of these

types of noise on conversation was originally the criteria on which units of measure were based. However, in order to disturb conversation, the noise must be reproduced in the handset. This meant that the frequency response of the handset played a part in determining what electricial noise was even reproduced. Therefore, the model numbers of some Western Electric handsets have accompanied the designations of some noise units (144, F1A). With the advent of modern data type communications, these older noise units which are based on speech are inadequate to express how disturbing a particular noise might be. The relatively straightforward 3 kHz flat weighting offers the most usefulness.

It has been found that sound power around 1000 Hz is very disturbing to a listener, since it interferes with his communication. A sound of equal power at 200 Hz or 5,000 Hz, injected into the listeners ear will be heard, but it will not disturb his ability to communicate nearly as much. These frequencies would have to be almost 25 dB stronger before the effect would be noticed by the listener to be the same as the 1000 Hz tone. This type of comparison has been made by using a Western Electric type 500 telephone set at all frequencies in the voice channel and each frequency given a weighting. Figure 6 shows the result which is called "C-Message" weighting. This curve might also be thought of as a filter response curve, since a filter is the device used to weight a noise reading. The curve shows that a tone at 200 Hz would have to be 25 dB higher in order to give the same power output from the filter network, as compared to a 1000 Hz tone.

With such a filter placed before the meter, the composite noise in the channel can be measured and read out in dB or dBm "C-Message" weighted. Other similar noise units include 144 line-weighting, F1A line-weighting,

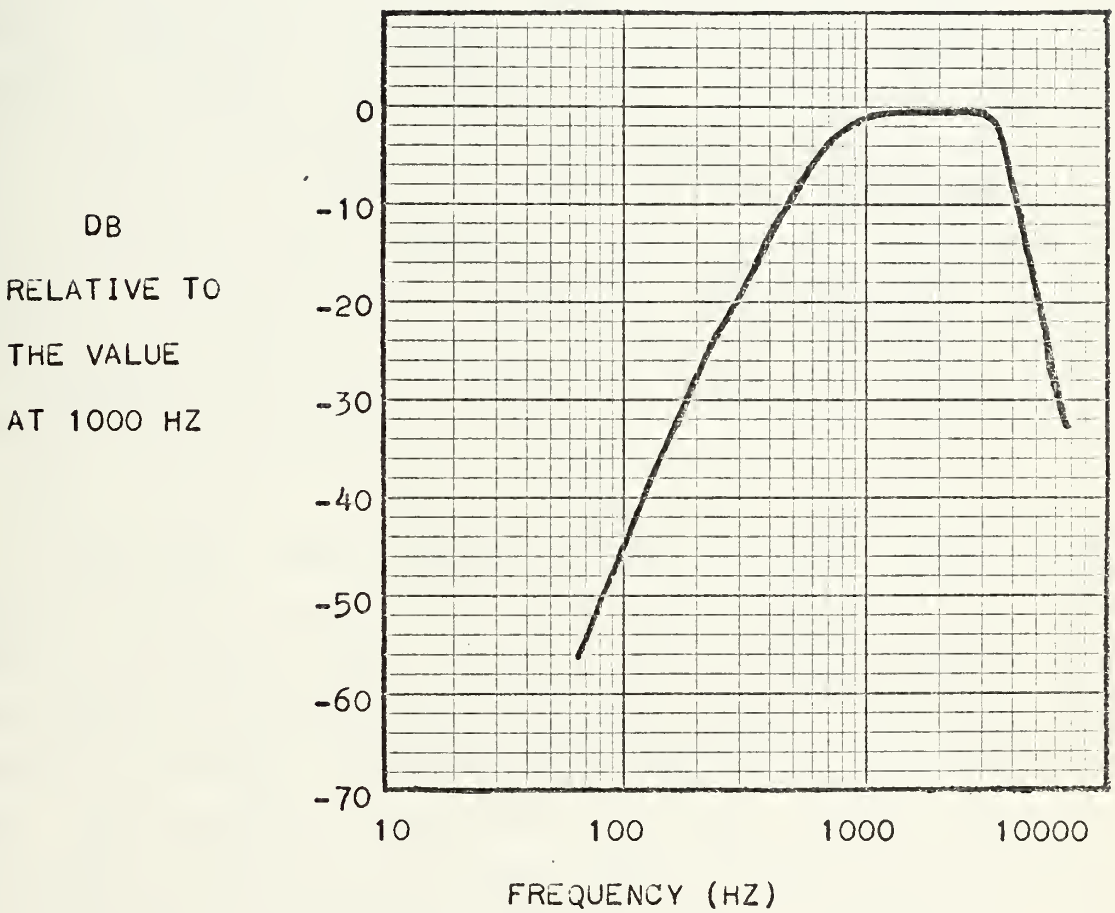


Figure 6.- C-Message Weighting Characteristic

psophometric-weighting, and flat-weighting. The flat-weighting curve is essentially one which allows from 0 to 3 kHz to be weighted evenly (no weighting) with a rapid cutoff beyond 3 kHz. This weighting is most nearly representative of all the noise in the voice channel weighted on an equal basis.

Once the weighting has been determined, the only other consideration is the unit for measurement (watts, dBm, etc.). One system involves measurement in picowatts and another involves dB related units. It was found that a 1 kHz tone had negligible disturbing effect on the human ear at a level of -90 dBm; therefore, -90 dBm is sometimes used as a reference level, allowing noise to be expressed in dBrnc (dB referenced to 1 picowatt, noise weighted, c-msg). Because the reference level is so low, all measured values are positive numbers. Figure 7 shows how such a noise reading might be taken.

Reference noise is then 1 picowatt (10^{-12} watt) or -90 dBm of 1000 Hz signal. The term dBrn refers to decibels above reference noise, which was the unit of measurement used with the orginal noise measuring sets. The term dBa refers to decibels above adjusted reference noise (1 dBa=-85 dBm). The purpose of this unit was that similar readings would be obtained on a noise signal consisting of thermal noise limited in frequency spectrum from 0 to 3000 Hz. The dBa as a noise unit is being supplanted by the dBrn. Picowatts of noise psophometrically weighted is pWp. Units of noise power derived from measurements with the CCITT recommened psophometer. This reference noise is -90 dBm at 800 Hz.

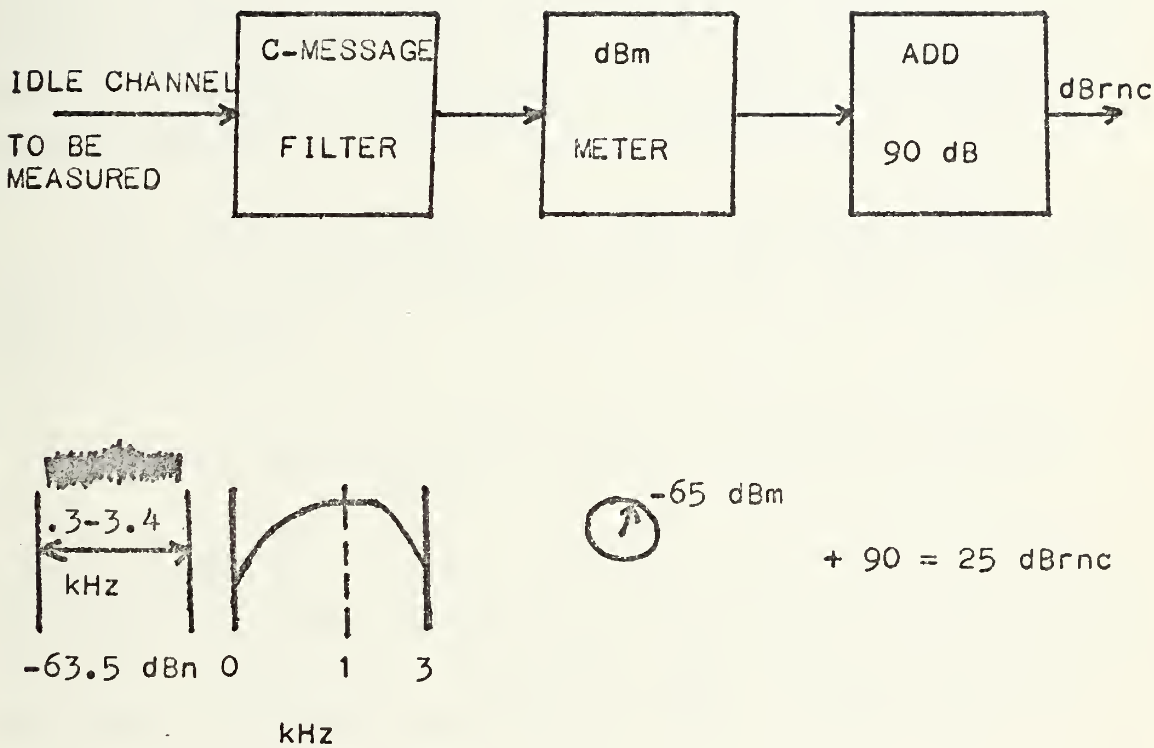


Figure 7- Idle Channel Noise Measured With A Weighting Filter (Computed in dBrnc)

Table 1 is helpful in correlating the various types of noise units with one another. It shows the relationship between five commonly used units for expressing noise in a voiceband channel. In the first four columns the units represent weighted noise at a point of zero relative level. In the fifth column the "S" represents a tone at zero relative level, and the "N" represents unweighted noise in a 3 kHz voice channel, therefore, S/N is the dB ratio of test tone to noise. This column may also be interpreted as negative dBm noise readings, flat weighted.

The table is based on the following commonly used correlation formulas, which include some slight round-offs for convenience.

$$\mathrm{dBrnc0} = 10 \log_{10} \mathrm{pWp0} = \mathrm{dBa0}+6 = \mathrm{dBm0p}+90 = 88-\mathrm{S/N}$$

Correlations for columns 2, 3 and 4 are valid for all types of noise. All other correlations are valid for white noise, but not necessarily for other types. The conversions are good only for evenly distributed noise. (Reference 19)

TABLE I

NOISE UNIT COMPARISON CHART

dBrnc0	dBa0	pWp0	dBm0p	S/N$_{dB}$	dBrnc0	dBa0	pWp0	dBm0p	S/N$_{dB}$
0	-6	1.0	-90	88	34	28	2520	-56	54
1	-5	1.3	-89	87	35	29	3162	-55	53
2	-4	1.6	-88	86	36	30	3981	-54	52
3	-3	2.0	-87	85	37	31	5012	-53	51
4	-2	2.5	-86	84	38	32	6310	-52	50
5	-1	3.2	-85	83	39	33	7943	-51	49
6	0	4.0	-84	82	40	34	10,000	-50	48
7	1	5.0	-83	81	41	35	12,589	-49	47
8	2	6.3	-82	80	42	36	15,849	-48	46
9	3	7.9	-81	79	43	37	19,953	-47	45
10	4	10.0	-80	78	44	38	25,199	-46	44
11	5	12.6	-79	77	45	39	31,623	-45	43
12	6	15.8	-78	76	46	40	39,810	-44	42
13	7	20.0	-77	75	47	41	50,119	-43	41
14	8	25.2	-76	74	48	42	63,096	-42	40
15	9	31.6	-75	73	49	43	79,433	-41	39
16	10	39.8	-74	72	50	44	100,000	-40	38
17	11	50.1	-73	71	51	45	125,890	-39	37
18	12	63.1	-72	70	52	46	158,490	-38	36
19	13	79.4	-71	69	53	47	199,530	-37	35
20	14	100	-70	68	54	48	251,990	-36	34
21	15	126	-69	67	55	49	316,230	-35	33
22	16	158	-68	66	56	50	398,100	-34	32
23	17	200	-67	65	57	51	501,190	-33	31
24	18	252	-66	64	58	52	630,960	-32	30
25	19	316	-65	63	59	53	794,330	-31	29
26	20	398	-64	62	60	54	1,000,000	-30	28
27	21	501	-63	61	61	55	1,258,900	-29	27
28	22	631	-62	60	62	56	1,584,900	-28	26
29	23	794	-61	59	63	57	1,995,300	-27	25
30	24	1000	-60	58	64	58	2,519,900	-26	24
31	25	1259	-59	57	65	59	3,162,300	-25	23
32	26	1585	-58	56	66	60	3,981,000	-24	22
33	27	1995	-57	55					

D• DISTORTION

Ideally, every amplifier, transmitter, transmission line, waveguide, or receiver should emit signals in exactly the same form as the input signals. The imperfection and inherent non-linearities of the physical materials used in these devices will not allow this. Therefore, the output signal is changed in shape from the imput.

A single frequency AC signal sent across a transmission line will travel at a particular speed, related to the type of material in the line and the frequency of the signal. Thus, the output waveform will not occur simultaneously with the imput, but will be delayed an amount of time depending on the length and electrical properties (capacitance, inductance) of the line as in figure 8. Such an effect is also true with any electronic device, only to a much lesser extent.

Because the line has resistance, the amplitude of the signal will be decreased or attenuated. The exact amount of attenuation is related directly to the type of material in the line and the environmental conditions around it. Therefore, the amplitude of the signal will be less at the output. Since only this one frequency is used, the delay and the attenuation will have no real effect on the shape of the signal or its usefulness, unless it is too small to be detected.

When a complicated signal of many frequencies (such as the human voice) is sent down the line, the effective inductance and capitance of the line will affect each frequency differently, slowing some up more than others and attenuating some more than others. The result is that the components of the signal will not be in the same order

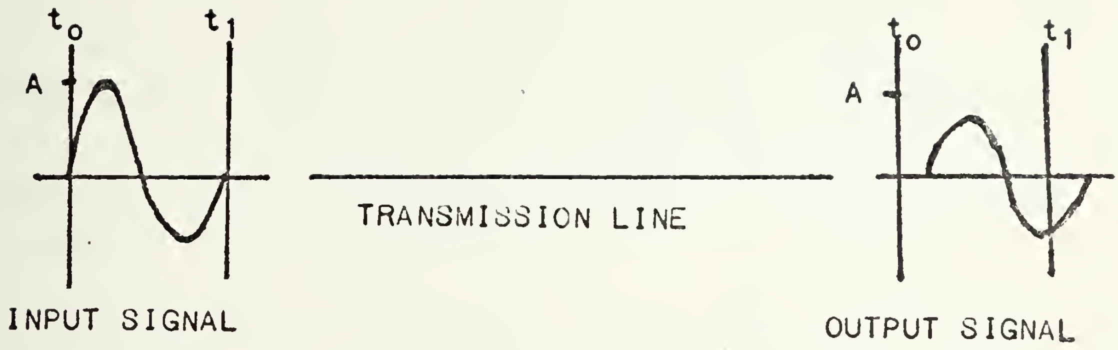


Figure 8- Attenuation and Delay of a Signal (Sinusoid) Down a Transmission Line

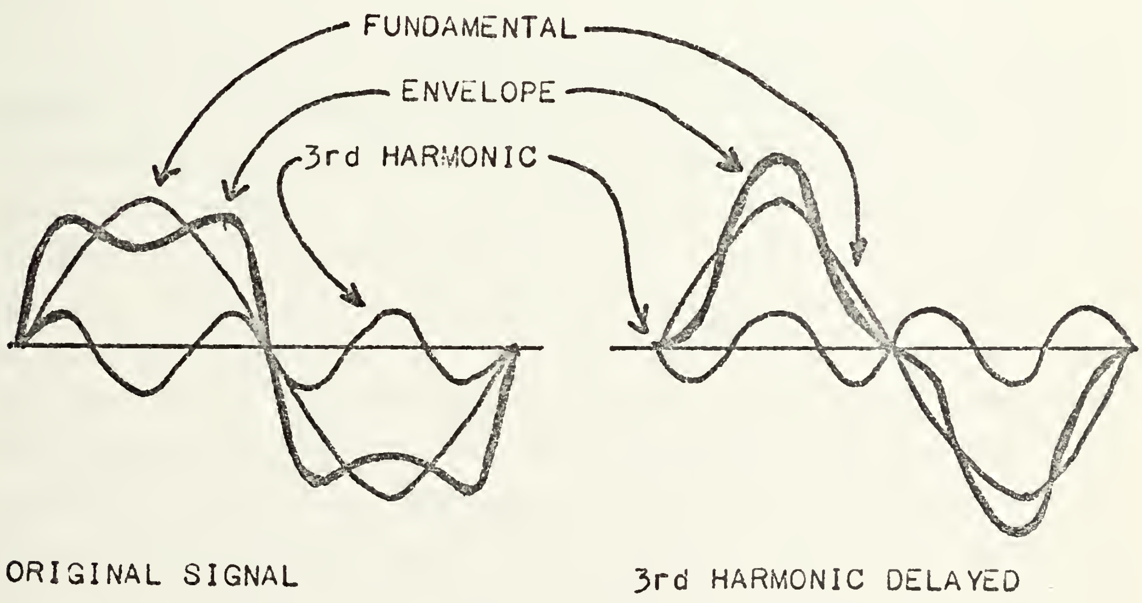


Figure 9- Effect of 3rd Harmonic Delay on Resulant Envelope of Signal

at the receiving end of the line. This can be more easily seen in figure 9 by an example using only two frequencies, a fundamental plus its third harmonic. If the third harmonic was delayed one-half cycle, while the fundamental was not delayed, the resultant waveform (algebraic sum) at the output would look significantly different from the input and might not be recognizable to detecting equipment. This is an example of delay distortion.

Phase delay is the result of the different amounts of phase shift over the frequency band (figure 10). The combination of these effects if called phase delay distortion, which produces the type of effect shown in figure 9. The important thing to remember is that the absolute phase delay is unimportant. It is the relative delay of the different frequencies of the signal that cause trouble.

The relationship between the phase shift (θ) and the frequency of a system (transmission line, amplifier, etc.) can be shown graphically. If the relationship is linear and the graph appears to intersect the zero degree frequency line at θ = 0 or 180 degrees, the delay is equal (or one-half cycle different) for all frequencies concerned and the output waveform is simply moved in time and not distorted as in figures 11A and 11B. If the phase-frequency relationship is linear, but the extended graph does not intersect the zero frequency line at 0 or 180 degrees (or multiples of 180 degrees), the signal will be distorted since the delay for each component frequency will be different. This effect is called intercept phase delay distortion and is shown in figure 11C. To determine the time delay of a single frequency, the following formula may be used :

$$t(\text{sec}) = \theta(\text{degrees}) / f(\text{Hz})\ 360\ (\text{degrees/cycle})$$

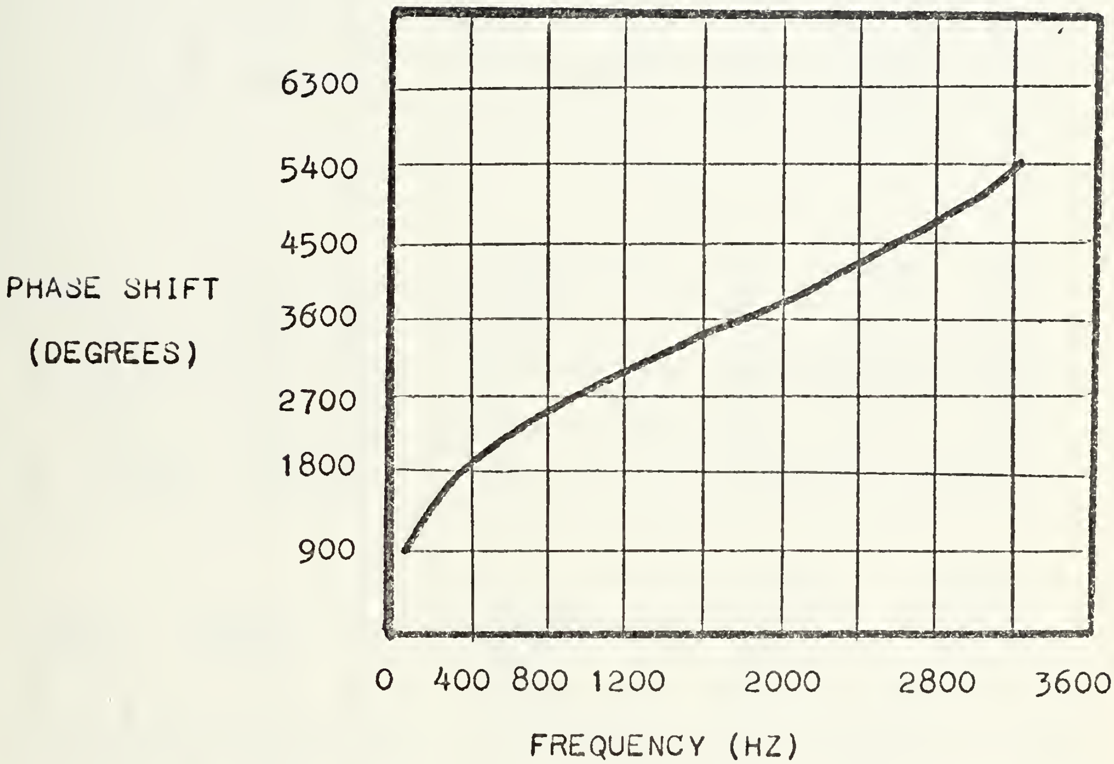


Figure 10- Phase Shift vs. Frequency
Typical phase shift characteristic of 100 mile carrier telephone circuit. Note that at no point is the curve straight.

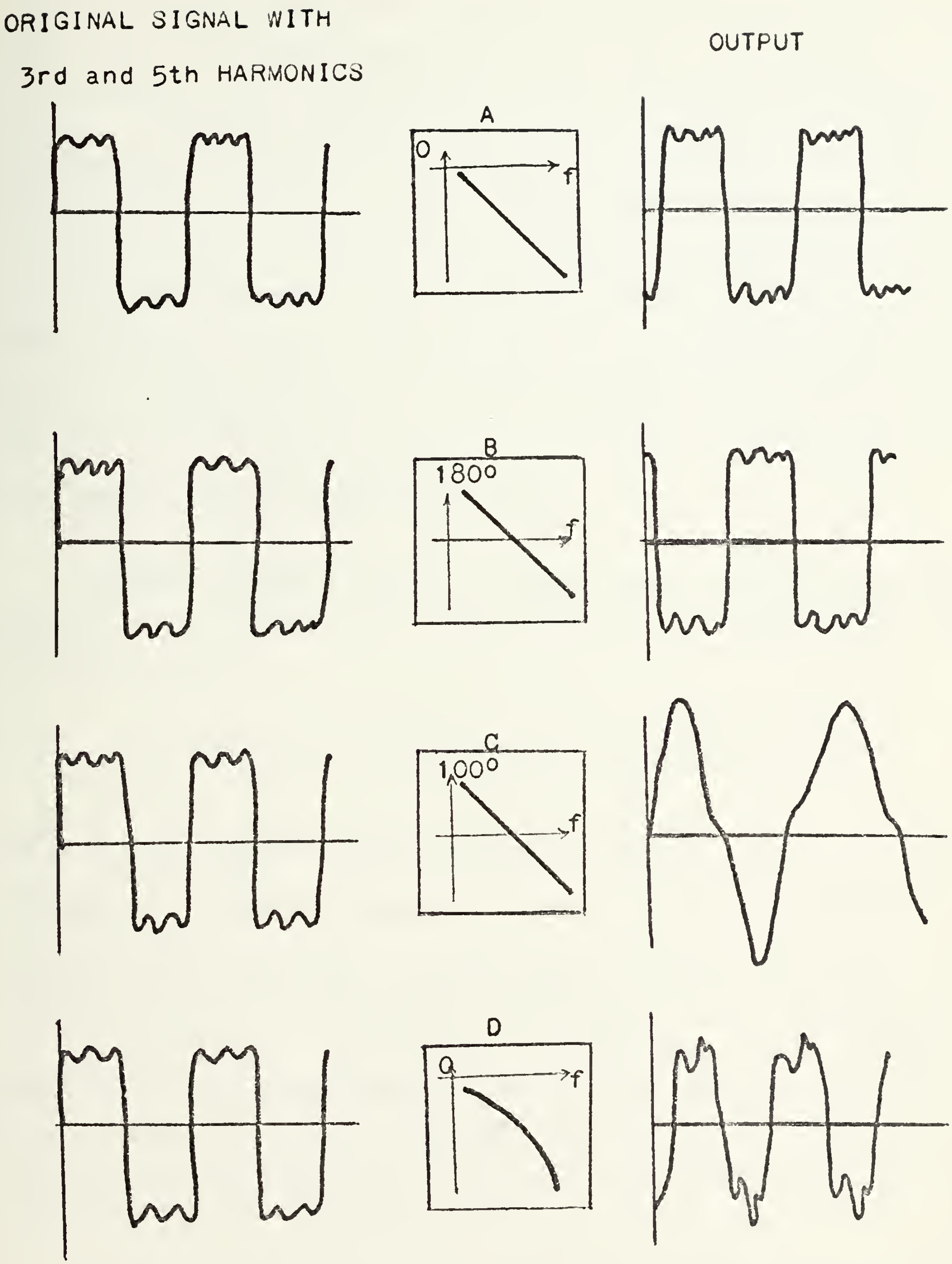


Figure 11- Effects of Various Phase-Frequency Relationships on a Complex Signal

For more complex waveforms such as amplitude modulation, the delay distortion characteristics are more complicated. In the simplest case, there would be one frequency modulated by another. Each of these frequencies would move down a transmission path at different velocities. The beat frequency which is the result of the other two would move at a third speed. If these velocities were greatly different, the resultant waveform would be noticeably distorted. This form of distortion, when applied to a complex modulated waveform, is commonly called envelope delay distortion. The amount of envelope delay is also dependent on the linearity of the phase-frequency relationship. This form of delay is very common in wire communications.

Envelope delay can be shown on a phase-frequency diagram as in Figure 12. The amount of envelope delay can be described as:

$$t(\text{sec}) = \theta_2 - \theta_1 (\text{degrees}) / f_m (\text{Hz}) 360 (\text{degrees/cycle})$$

Where f_m is the modulating frequency and $\theta_2 - \theta_1$ is the difference in phase shift between the carrier frequency and the sum or difference frequencies. The result of the envelope delay action shown in figure 13. The absolute is all that is shown. This effect can also be called sideband delay and is applicable to any form of modulation. Other common names for envelope delay are "relative delay" and "group delay". This is usually referenced to a middle frequency in the bandwidth.

Delay distortion is a very real trouble for digital pulse transmission. When a signal switches from a mark (1) to a space (0) or vice-versa, the switching action generates

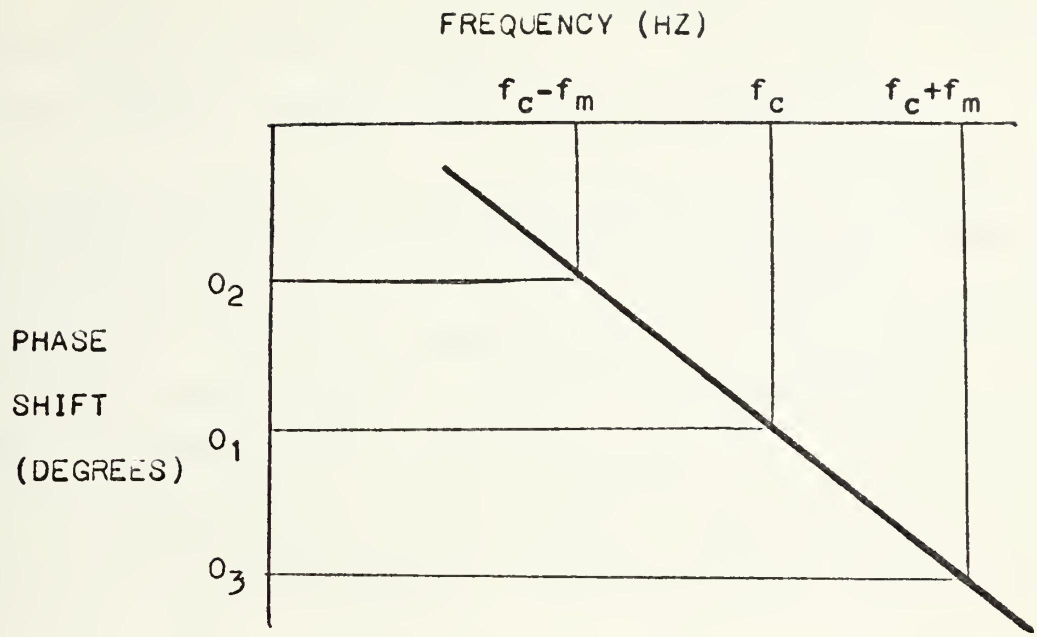


Figure 12- Phase Shift vs. Frequency Chart

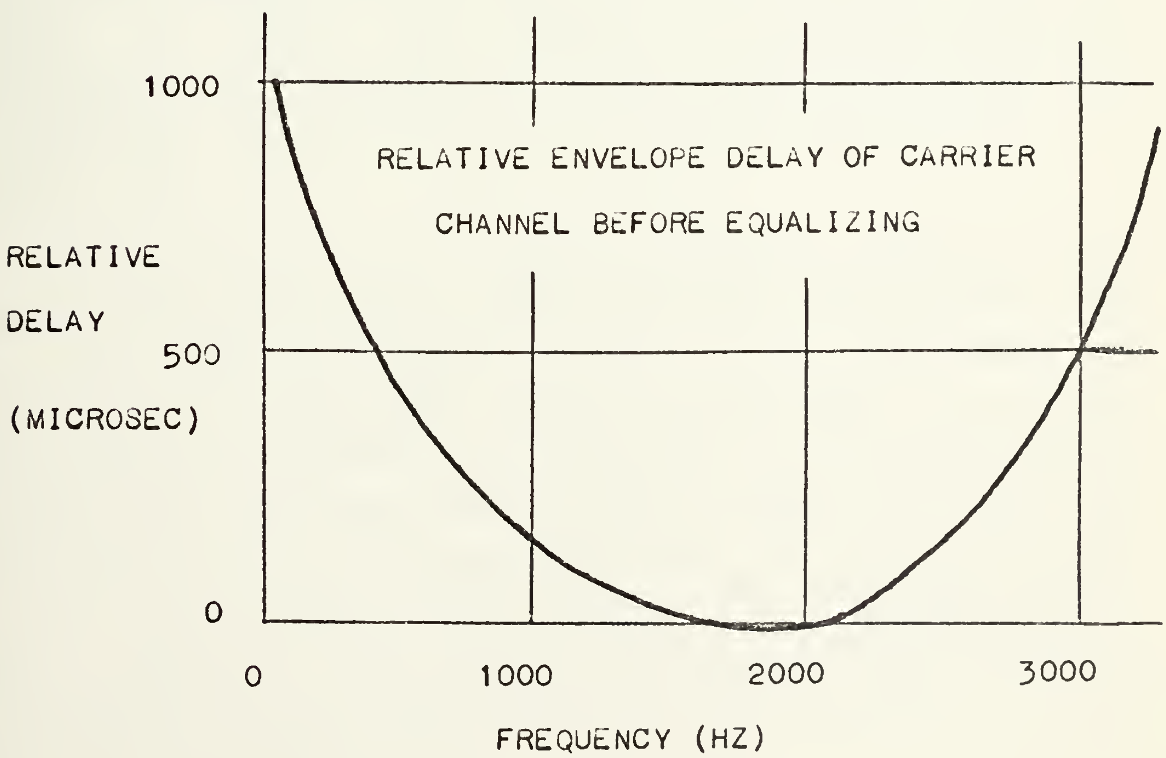


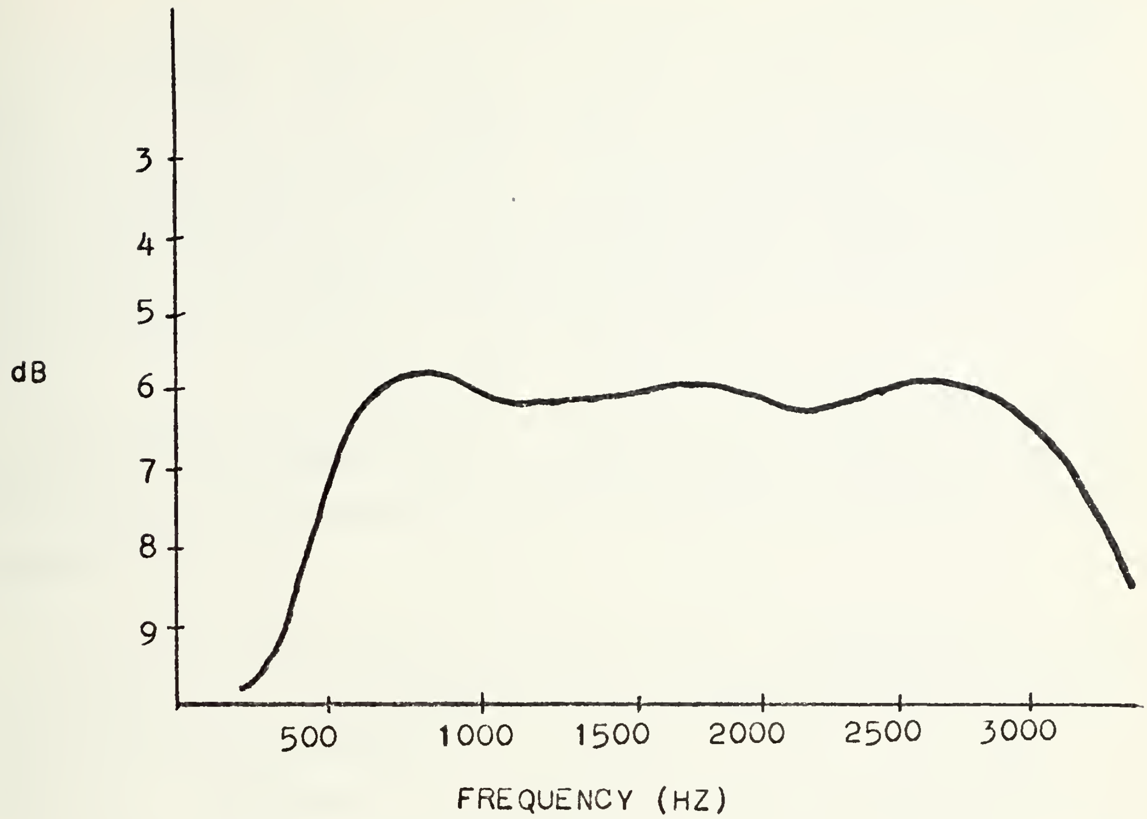
Figure 13- Typical Envelope Delay

a wide variety of frequencies simultaneously and the delay distortion will seriously distort the pulse train. With high speed digital circuits this can be very critical.

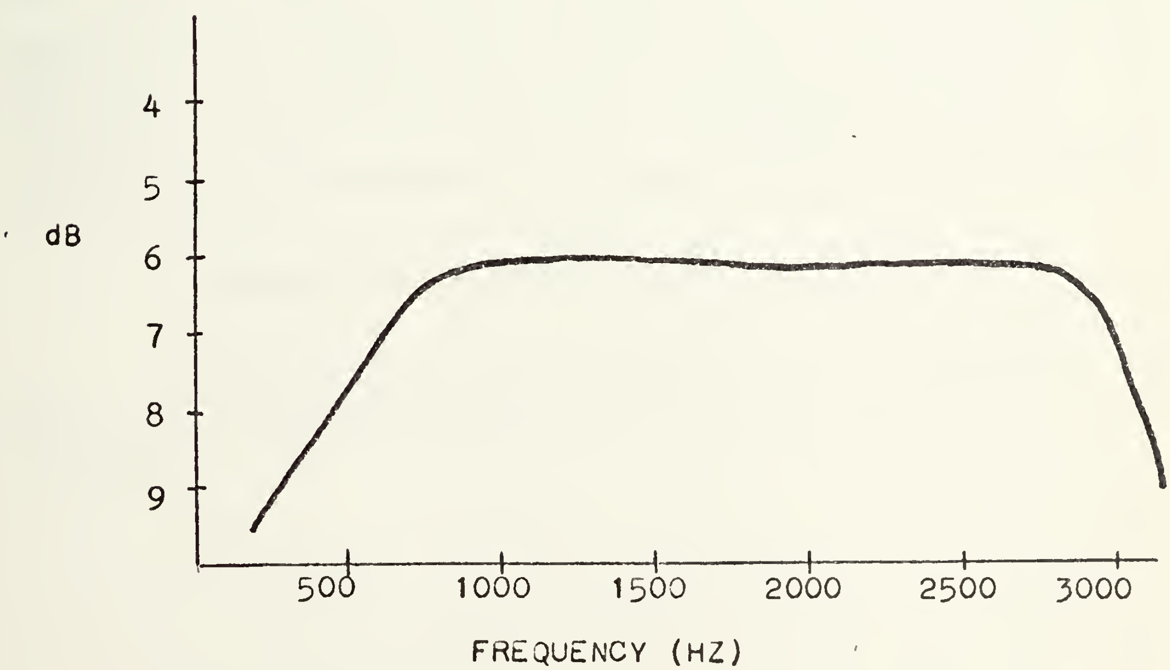
A frequent error is to confuse true delay distortion, as determined from the phase characteristic, with envelope delay distortion, which is obtained from the envelope delay characteristic. Part of this confusion is the result of nomenclature adopted years ago for telephotograph and television transmission. As then defined, envelope delay distortion (EDD) was the maximum deviation in envelope delay across a certain band of frequencies. Delay distortion is related to, but not the same as, envelope delay distortion.

Frequency response or amplitude distortion refers to the differences in attenuation for different frequencies in the bandwidth of the signal. This can occur in conjunction with phase distortion but is not necessarily part of it, figure 14 shows two examples of frequency response distortion. The distortion is usually rather irregular, but a smoother response can often be approximated as shown in figure 14B.

This distortion is often caused by filters in the system, which produce band edge roll-off. It can also come in the form of inband ripple, often the result of reflections and echo from impedance mismatches in the line. The band edge roll-off is of importance in low speed digital traffic where frequency shift keying is used. The band edge distortion can cause signal amplitude for those frequencies to fall below recognizable levels. For high-speed data, delay distortion does much of the damage and only impedance mismatches from switching points have major effects on the amplitude characteristics.



A. Actual Frequency Response



B. Idealized Frequency Response

Figure 14- Frequency Response Distortion

Echo is named after the audible effect of a returned sound interfering with originally produced sound. It is the return of part of a signal back up the transmission line over which it was propagated. It is especially a problem in voice communications and often results from hybrid circuits. Two-to-four wire hybrids are not perfectly balanced and some of the signal will return down the line.

Echo and reflection are related in that they cause basically the same effect. Reflection is usually referred to in general transmission line discussions as echo in voice transmissions.

Impedance mismatches or faults in a transmission line can bring about reflections of the signal which can interact with the incoming signal. An example of this is shown in figure 15. Depending on the amplitude and relative phase of the reflected wave, the interaction with the incident wave can cause phase and amplitude-like distortions. The distortions are directly proportional to the reflection coefficient of the line and the modulating frequency. The reflection coefficient is is calculated from the characteristic impedance of the line and its load impedance:

$$\text{Reflection Coefficient} = (Z_L - Z_o)/(Z_L + Z_o)$$

Where Z_L is the load impedance and Z_o is the characteristic impedance.

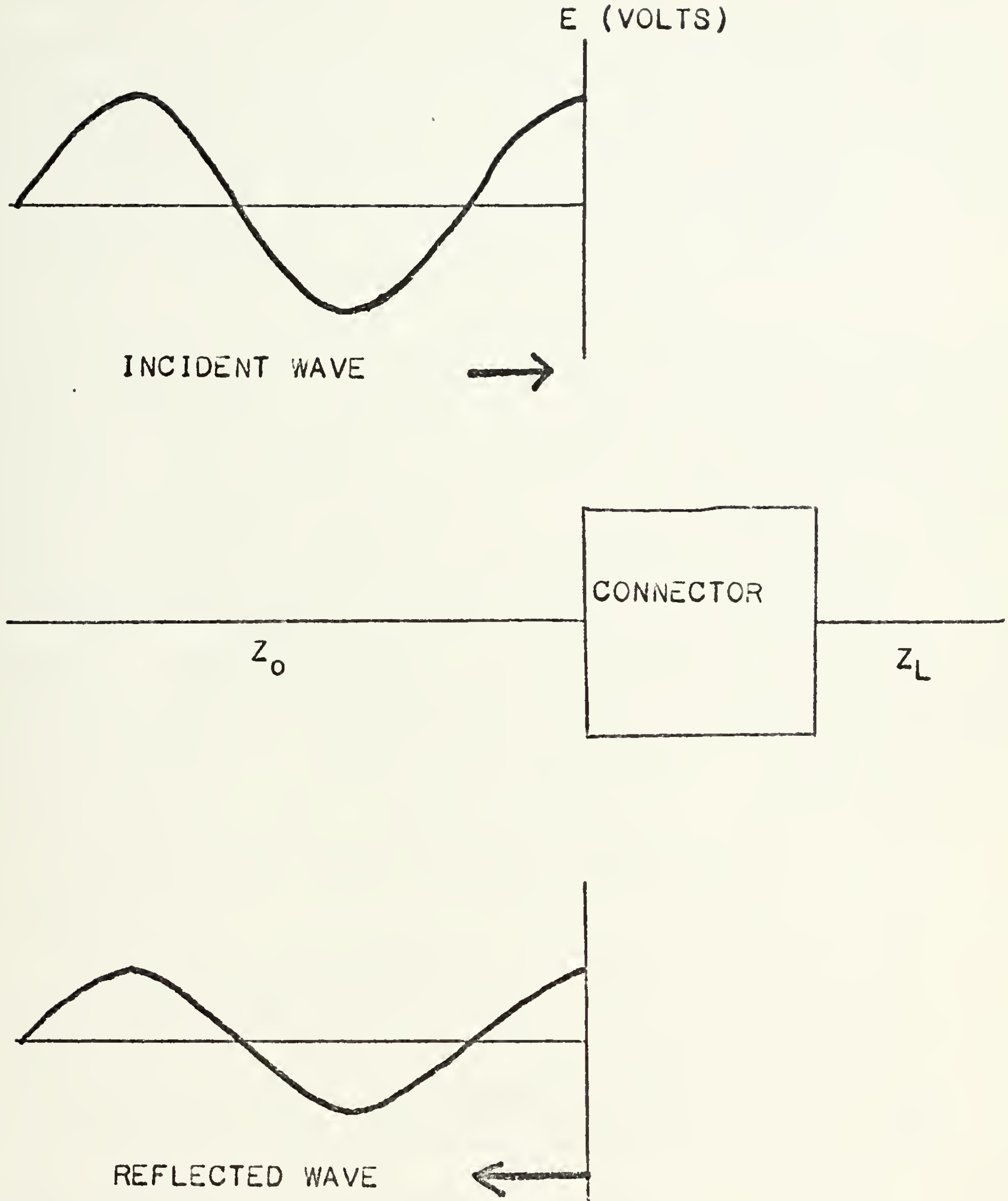


Figure 15- Reflection of an Incident Wave by a Mismatched Impedance

Imperfect impedance matches at switch points, transformers, and device imputs can affect the reflection coefficient to the point that serious phase distortion could result from the reflections. The resultant phase distortion is of the same type as described as delay distortion except that here it is the result of the interaction of the original and reflected signals on the line.

Two forms of distortion that can occur when a signal consisting of many frequencies (such as AM or FM) is transmitted are harmonic and inter-modulation distortion. Whenever two or more frequencies are tramsmitted together through a nonlinear device, they interact with each other to form frequencies equal to the sum and difference of the originals. Nonlinearities in the transmission media also tend to give rise to multiples of the original frequencies called harmonics. These two phenomena occur in various ascending orders. For example, given two frequencies, f_1 and f_2, the second order results would be:

$2f_1$, $2f_2$ (harmonics)

$f_1 + f_2$, $f_1 - f_2$ (intermodulation)

Likewise, third order results would be :

$3f_1$, $3f_2$ (harmonics)

$2f_1+f_2$, $2f_1-f_2$, f_1+2f_2, ect. (intermodulation)

The higher orders would be in similar forms.

Harmonic distortion can be caused by echo or by nonlinear operation. Harmonic distortion is measured by taking the ratio of the magnitude of the harmonic to the magnitude of the original signal. It is usually expressed as a percent :

$$\text{Harmonic Distortion} = (A_2 / A_1) \times 100\%$$

Where A_2 is the magnitude of the second order harmonic and A_1 is the magnitude of the fundamental. Total harmonic distortion is measured by :

$$\text{Total Harmonic Distortion} = \sqrt{(A_2^2 + A_3^2 + A_4^2 + \ldots) / (A_1^2)} \times 100\%$$

Where A_2, A_3, ect. are the magnitudes in voltage of the various harmonics.

Intermodulation distortion is caused by much the same effect as harmonics, often in the modulator section of the system. The magnitude of the modulating signal controls the intermodulation amplitude. The amount is again expressed as a percent or dB ratio to the original levels. Although harmonics often appear with intermodulation distortion they are not considered part of it.

A problem which has not received much attention before now is that of phase jitter, which is undesired phase modulation of received data signals. The instantaneous phase of the received data signal is likely to jitter, typically, at rates of 180 Hz and below causing sidebands with magnitudes of approximately 18 dB below the level of the carrier. This is approximately 15 degrees peak-to-peak.

This effect is primarily caused by ripple in the DC power supply appearing in the master oscillator of long haul carriers and being multiplied through many stages. Some phase jitter also occurs in short haul systems from incomplete filtering of image sidebands. Digital carrier systems also will exhibit jitter at certain input frequencies. Figure 16 is an example of what phase jitter of a single sinusoid would look like on a oscilloscope. Jitter is shown by the "smear" of the signal over the horizontal axis and is measured in degrees.

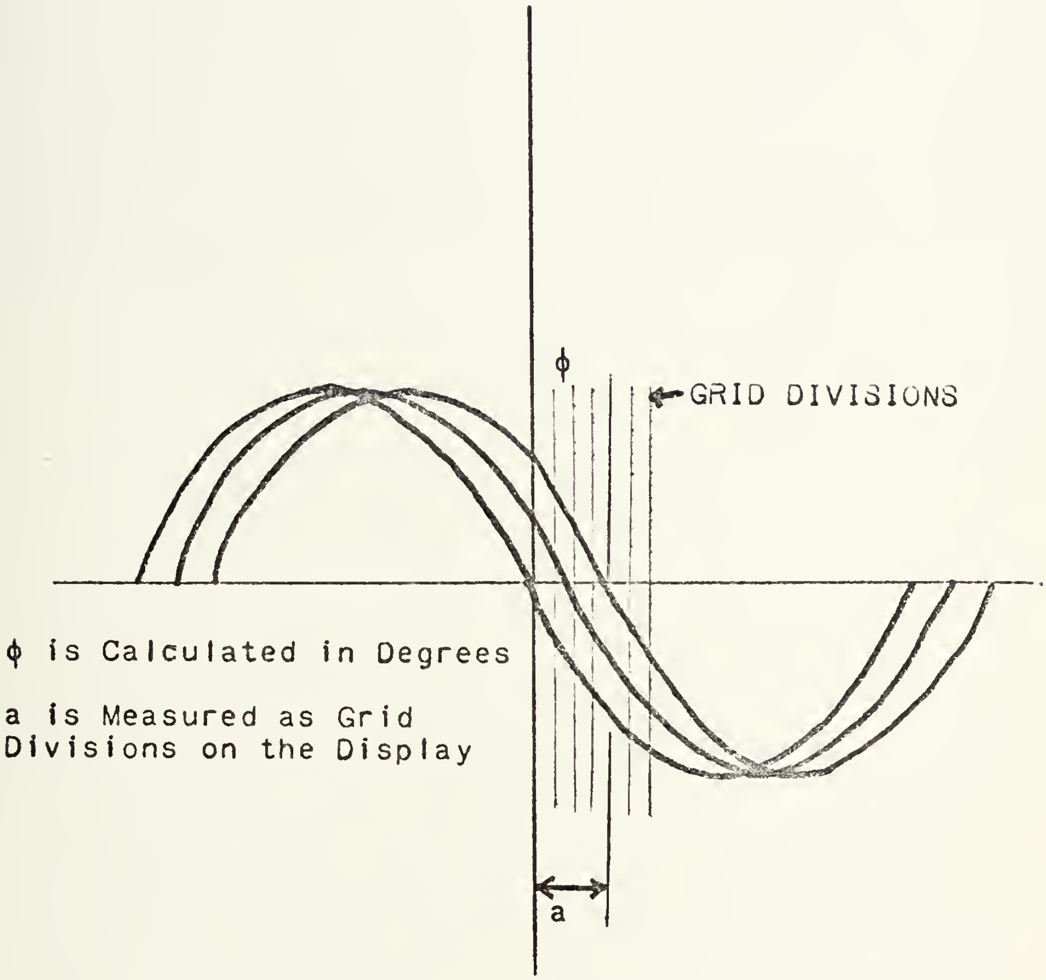


Figure 16- Oscilloscope Display of Phase Jitter

E. SUMMARY

It is evident from the discussion that the three areas of consideration are levels, noise, and distortion. It is also evident that there is a great deal of interaction between these areas. The key indicators were discussed in general terms as to their relationship to disturbances that may occur in communications circuits. The next step then is to consider each of these key areas in detail and determine the specific, measurable parameters that can be analyzed (manually or electronically) to allow adequate prediction of system failure.

IV. PARAMETERS REQUIRED FOR PERFORMANCE MONITORING AND ASSESSMENT

A. LEVEL

The level characteristics of a transmission medium are of fundamental importance. In contrast to a linear passive network, these quantities are not always readily measureable on a real transmission facility. Furthermore, these characteristics may be signal dependent to equipment such as compandors. Syllabic compandors used in analog facilities introduce loss during quiet intervals of speech to reduce noise and subjectively improve the channel for speech. Instantaneous compandors are used in digital facilities to provide an approximately constant signal-to-noise ratio, independent of the signal power over the design range of the facility. (Reference 20)

1. Loss

The net loss of a communication channel must be known so that correct operating levels can be established and maintained. For most communications channels the net loss or gain should not change significantly from hour-to-hour or from day-to-day. It should also not change significantly when the input level is varied within the operating level range of the channel. Limits are established for the amount of variation allowable in the net loss over a period of time. Limits are not usually established for the variation in the net loss or gain of channels with variation of input level, but any significant variation in the net loss when the level is varied within the operating range indicates the channel is not operating properly.

The loss of a channel, expressed in dB, is the loss

experienced by a single frequency tone in traveling the transmission medium. The power of the received tone is measured on an averaging type of instrument, calibrated in dBm. Due to mistracking in some compandored facilities, the loss may vary as the test tone power level is changed.

2. Return Loss

Impedance mismatches in a circuit give rise to reflections of signal energy called echoes. A single reflection causes energy to return to the transmitter (talker echo). If the single reflection is again reflected at an impedance mismatch, signal energy will arrive at the receiver some time after the original signal (listener echo). Echoes due to multiple reflections also occur, but they are usually insignificant by comparsion with those due to single or double reflections.

Echoes generally cause no problems on 4-wire channels. However, on 2-wire channels, and on channels containing both 2-wire and 4-wire sections, echoes may be interfering to both voice and data communications. Echoes are controlled by matching impedances and by controlling loss in the echo path. Figure 17 illustrates the use of a hybrid transformer which joins 2-wire and 4-wire facilities. The impedance of the balancing network ideally is the same as the impedance looking into the 2-wire facility. If this condition is met, signal power from the receive side to the 4-wire facility will divide equally between the balancing network and the 2-wire facility and no signal power will "return" on the transmit side of the 4-wire facility. In practice, the match is never perfect. Some portion of the received signal power is thus returned on the transmit side of the 4-wire facility. Various parameters are used to determine channel quality with respect to echoes. These include return loss, echo return loss, singing return loss, and singing margin.

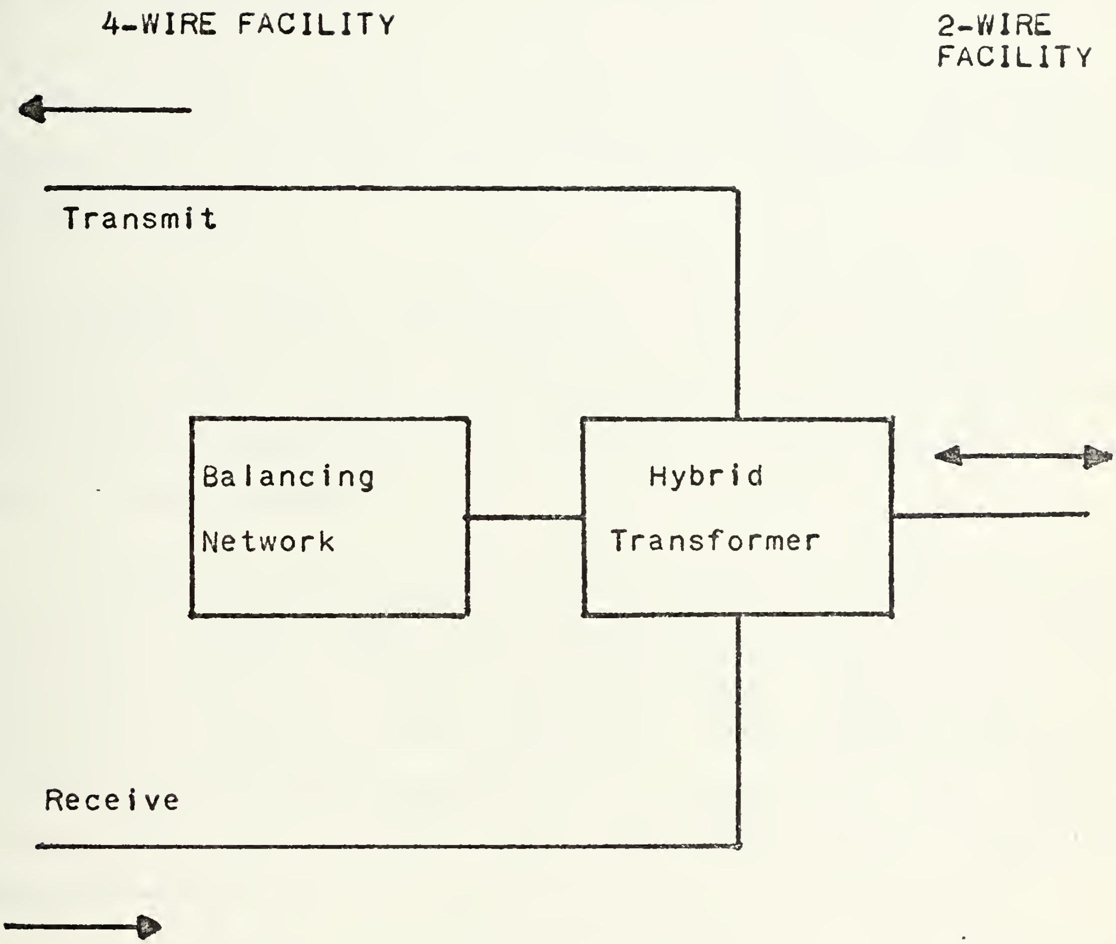


Figure 17- Typical Interface Between 2-Wire and 4-Wire Facilities

Return loss is the ratio in dB of the power of a single frequency signal placed on the receive side of the 4-wire facility to the resulting power at that frequency appearing on the transmit side.

Echo return loss is a weighted average of the power returned at all frequencies between 500 and 2500 Hz and expressed in dB.

Singing return loss is also a weighted average of the reflected power at all frequencies in a frequency band expressed in dB. There is a low frequency test covering the 200 to 500 Hz band, and a high frequency test covering the 2500 to 3200 Hz band.

A singing point test is made by placing a variable gain amplifier between the two sides of the 4-wire facility and increasing the gain until self-sustained oscillation, or "singing", occurs at some frequency. The singing condition is detected by audio monitoring of the circuit. The amount of gain in dB required to cause singing is called the singing point.

A singing margin test is similiar to the singing point test. In this case, the channel is terminated in its normal impedances. A variable gain amplifier is inserted on one side of the 4-wire facility and the gain increased until singing occurs. The amount of gain in dB required to cause singing is called the singing margin.

3. <u>Long Term Loss Variation</u>

Due to normal aging and drift of amplifiers, temperature changes in cables, and changes in the physical makeup of a channel, changes in the loss of a circuit can be

expected. Such changes are commonly referred to as seasonal changes, emphasizing the effects of temperature.

Routine tests are made on transmission facilities between installations to maintain loss deviations within nominal values. Thus the greatest changes, from the user's point of view, normally come about due to temperature changes on the cable connecting the subscriber to his central office. No special test equipment, other than that used for frequency response or simple loss measurements, is required to detect these changes.

4. Attenuation Distortion

The amplitude characteristic of a network or system is commonly determined by simply measuring the loss of a single frequency tone as it is tuned across the bandwidth of interest. This measurement is referred to as the static frequency response. On telecommunications channels, due to level and frequency sensitive devices, and the presence of nonlinearities, a static measurement may not yield the same amplitude characteristic as that experienced by a complex waveform. Difference of up to 2 dB have been observed on some compandored facilities when using other measuring techniques. These differences arise from the relative placement of filters with respect to level sensitive and nonlinear devices and generally are greatest near the band edges. In these cases, the frequency response is a function of the spectral content of the signal on the channel. This latter response is referred to as the dynamic frequency response. Comparison tests have shown that the dynamic response can be measured by using two tones. One tone, in the vicinity of 1000 Hz is referred to as a holding tone, and serves to activate the channel. The second, or measuring tone, is used to measure the frequency response. The holding tone is held 5 dB above the measuring tone and

the response is measured using a frequency selective voltmeter. The composite test tone power should approximate the expected signal power that will appear on the channel when in normal use.

Even though differences (+2 dB) may be encountered between the static and dynamic testing methods, the static one has long been the method of measurement in military telecommunications. Because of its simplicity, it will probably continue to be used. The differences encountered on analog companded facilities are generally in a direction such as to make the channel appear to have a narrower bandwidth than it does in the presence of a signal with a broad spectrum.

Frequency response can be checked and controlled by the following simple test: the loss on a channel is measured at three different frequencies, 400 Hz, 1000 Hz, and 2800 Hz. The loss at 1000 Hz is then subtracted from the loss at 400 and 2800 Hz. These differential losses are then referred to as the slope at 400 or 2800 Hz. These two slopes are a measure of the frequency response of the channel under test.

5. Bandwidth

The bandwidth of a channel is determined from measurements made of the amplitude characteristic. It is defined as the band of frequencies within which the loss is no more than 10 dB greater than the loss at 1000 Hz. Experience has shown that the simple slope tests are sufficient to control the bandwidth on transmission facilities in use. Therefore, bandwidth measurements are seldom made.

6. Baseband Loading

In radio and multiplex equipment, any high levels in the baseband raise the loading level to the modulator and can cause severe intermodulation noise in all channels. (Figure 3)

Multiplexing produces a composite output signal called the baseband. This composite signal contains group, super-group, pilots, signaling frequency (SF) tones, voice, teletype, data signals, and special steering signals produced by networks such as AUTOVON and AUTOSEVOCOM. When all these signals are combined into a baseband, they form a complex modulating signal.

In a multiplex system, even where the transmission level of each channel is fixed, the level of the composite signal varies. Since many different frequencies are transmitted, the phase relationship between these various frequencies varies randomly. Sometimes several frequencies will reach a peak together, causing a momentary rise in total power. At other moments the various frequencies may combine to lower the total signal power well below average. Since numerous signals are applied to a common amplifier in a multiplex system, the phase combinations that may occur greatly increase. There is always the chance that many or all of the signals will achieve a peak value simultaneously. However, as the number of signals increases, the chance they will all reach a peak value simultaneously decreases.

Speech signals are more complicated than data signals. Each signal consists of a variety of frequencies and a great range of amplitudes. Since amplifiers, modulators, and the like are responsive to instantaneous peak signal values rather than average or RMS values, multiplex systems must be

adjusted to accommodate the possibility of large peak values. In normal practice, the total average power level to the carrier system is set so that the overload or break point is not reached more than 1 per cent of the busiest hour. Since the power distribution varies with the number of channels on the system, the amount the average signal level must be reduced (the peak factor) will vary with the size of the system. Figure 18 shows peak factors for speech and tones in terms of the number of active channels. (Reference 19)

Baseband loading is the measure of the average (RMS) amplitude of this modulating signal. The signals are of three categories:

a. Constant level-constant frequency signals such as pilots, signal frequencies or special non-varying steering signals.

b. Constant level-varying frequency signals such as voice frequency carrier telegraph, data modem outputs, AUTOVON or AUTOSEVOCOM.

c. Complex signals with varying amplitude and frequencies such as speech or voice.

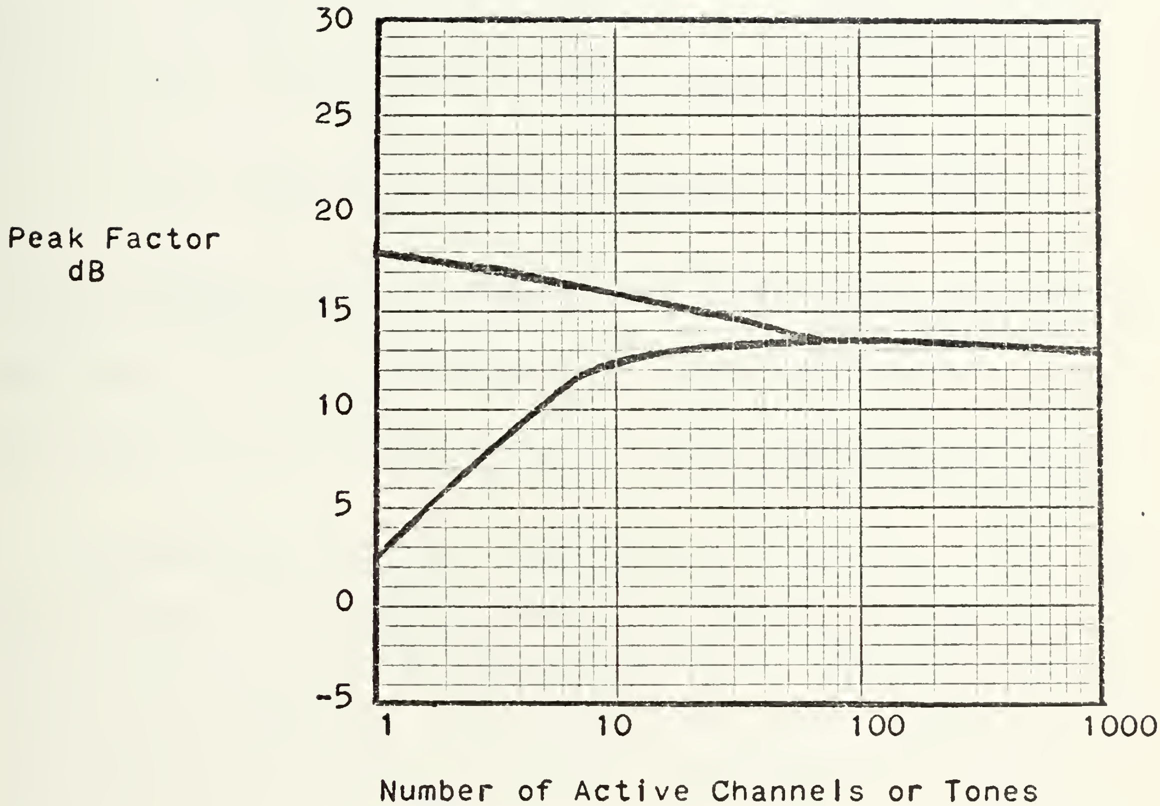


Figure 18- Multichannel Peak Factor. Values of Signal Peaks (above RMS) Exceeded Only 1% of the Busiest Hour.

The baseband loading of a multiplexer with constant level signals can be computed simply by the following formula:

$$P_T = P_i + 10 \log_{10} n$$

where P_T = total RMS power in dBm (or dBm0)

P_i = equal level of each channel

n = the total number of channels

If the signals are not of equal amplitude then conversion is made from dBm to mw (mw = $\log_{10}^{-1}$ (dBm/10)) then are added. The sum in MW is converted back to dBm (dBm=10 $\log_{10} P_{mw}$).

Speech can only be statistically approximated. The statistical analysis of speech has resulted in defining the average speech level as being between -16 dBm to -8 dBm. Several mathematical equations have been adopted by the International Committee for Radio Communications (CCIR) to determine the RMS equivalent load factor for multichannel system. These equations are referred to as the CCIR load equations:

$$P = -1 + 4 \log_{10} N \qquad (N \leq 240 \text{ channels})$$

$$P = -15 + 10 \text{ lob}_{10} N \qquad (N \geq 240 \text{ channels})$$

The -1 and -15 allow for the fact that not all voice channels are busy at the same time. They also make

allowances for multiplex pilots and SF tones. They do not allow for data or teletype signals. These constant amplitude signals must be calculated separately and their composite load level added to the CCIR predicition. (Reference 21)

7. Summary

In summary the parameters that are important in regards to level are:

a. Loss: End-to-end circuit attenuation, usually measured at 1000 Hz.

b. Return Loss: A measure of the mismatch between the actual circuit impedance as compared to a nominal impedance defined for ideal circuits.

c. Long Term Loss Variations: Changes in the loss of a circuit due to aging of components, changes in physical makeup, and temperature variations.

d. Attenuation Distortion: Loss deviation (1000 Hz reference) over the range of frequencies of interest. This includes static and dynamic frequency response and bandwidth.

e. Baseband Loading: The result of high levels in the baseband of multiplexed circuits. These high levels in the baseband raise the loading level to the modulator and can cause severe intermodulation noise on the channels.

B. NOISE

Noise on telecommunications channels arises from numerons sources. Some of these are thermal noise from amplifiers, 60 Hz and its harmonics which may be picked up from power line induction, intelligible or unintelligible crosstalk, single frequency tones, and switching and signaling transients. All of these effects and others come under the general heading of noise. Background noise is referred to as white noise, gaussian noise, message circuit noise, ect. What is intended is a frequency weighted measure of the total power on a channel not arising from the desired signal.

1. Signal Uncorrelated Interference

Primarily because of the existence of syllabic compandors and digital transmission facilities which require quantizing, two types of background noise must be distinguished. The first is called C-message noise and is the total frequency weighted noise power measured on a channel in the absence of signal. Hence, it is signal uncorrelated interference.

The second is referred to as C-message notched noise (c-notched noise) and is a measure of unwanted power in the presence of a signal. When signal-to-noise ratios are discussed, C-notched noise is usually implied. Quantizing noise, inherent in PCM systems is measured as C-notched noise. (Reference 22)

C-message noise is measured by a quasi-RMS type of instrument with a time constant of about 200 milliseconds. The noise is measured through a filter called C-message weighting which shapes the noise in such a fashion as to make the measurement more meaningful in terms of anncyance

to people listening to the noise with a telephone receiver. The C-message weighting characteristic was shown in Figure 6. The characteristic is relatively flat over most of the frequency range usually of concern for data transmission (600-3000 Hz). Thus this weighting is useful for data channels even though it was developed for voice applications. The accuracy of noise measuring sets of this type is typically ±1.0 dB as determined by its measurement of a known power source of 1000 Hz tone. (Reference 23)

C-message noise and C-notched noise readings are expressed in units of dBrnC, or dB with respect to reference noise (-90 dBm, or -90 dB with respect to 1 milliwatt). The "C" refers to the C-weighting. For example, C-weighted noise having an rms power of -70 dBm would be expressed as 20 dBrnC (90-70=20).

C-notched noise is measured with the same type of instrument as message circuit noise except for the filter characteristic. In order to estimate signal-to-noise ratios, a 1004 Hz tone (called a "holding tone") is applied at the far end of the channel at a power approximating an actual signal. This tone activates compandors and quantizers. The tone is removed at the receiving noise measuring set by a notch filter which suppresses the holding tone by at least 50 dB. A frequency of 1004 Hz is used instead of 1000 in order to avoid using a rational submultiple of the sampling frequency which can lead to problems.

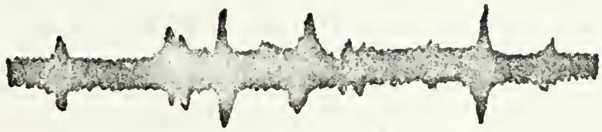
When either message circuit or C-notched noise is measured, small swings of the meter (up to ±6 dB) are mentally averaged. Occasional large jumps in the noise are ignored. These momentary large fluctuations are usually due to impluse noise which is measured in a different manner.

2. Impulse Noise

Impulse noise can enter a channel at any point and be carried throughout the system. Impulses can be generated in the equipment itself, be picked up from power lines, or be generated in the environment and radiated into the system. Excessive impulse noise can severely impair data transmission and must be traced to its source when it is detected. (Reference 24)

Impulse noise is a distinct phenomena from message circuit noise. Although some peaks of message circuit noise are registered on impulse noise counters, the vast majority of impulse noise arises from sources that are independent of message circuit noise sources. This case is demonstrated in Figure 19. The difference between message circuit noise and impulse noise is easily seen. In this case, the impulse noise is so apparent because of the relatively large bandwidth involved. Figure 19b shows the result in a voice bandwidth channel of one of the impulses illustrated in Figure 19a. Note the smear in time (5 milliseconds full scale) and the ringing due to the baseband filters. It is noise of the type shown in Figure 19b that is of primary concern.

The most frustrating characteristic of impulse noise is its time variability. When exposed to impulse noise measurements for the first time, the initial reaction may be uncertainty and lack of faith in any single measurement made in a reasonable short (5 to 15 minutes) time. This is primarily due to the fact that the number of impulses occurring during a fixed time interval at a fixed threshold is log-normally distributed. A typical sample of six such numbers might be 5,8,52,6,6,4. It is somewhat simpler to look at impulse noise that is normally distributed, thus that will be discussed first. (References 25 and 26)



A. 50 ms of Noise on a 48KHz Channel

B. 5 ms of Impulse Noise in a Voiceband Channel

Figure 19- Examples of Impulse Noise on Two Different Bandwidth Channels

Impulse noise level is defined as the threshold (expressed in dBrnC) at which the median count from a number of observations (each having the same specified time interval) is equal to a specified number. Both the median number of counts and the specified time interval have changed over the years. In the mid-1950s, the number of counts (not exactly the median in this case) was 70 and the specified time interval was 1 hour. In recognition of the fact that 1 hour is inordinately long for measuring a parameter, the test interval has been steadily shortened as more knowledge has become available.

By 1963, the measurement interval had been reduced to 30 minutes solely on the basis of a great deal of experience by numerous people who agreed that such an interval was adequate. On that basis, the 1963 impulse noise survey was conducted by making 30 minute tape recordings of noise. The data collected continues to be the best source of information on this topic. The data were summarized in two ways that are pertinent for this thesis. (Reference 27)

First, cumulative distributuion functions (cdf's) of the noise counts on individual channels were made every 5 minutes from 5 to 30. Two examples are shown in Figures 20 and 21. The impulse noise level may be tracked over the 30-minute interval by observing the threshold at which five counts in 5 minutes occurred, ten in 10 minutes, etc. (Impulse noise level is here defined as that threshold at which the median count for a number of observations is equal to one per minute.)

Traces of the impulse noise level may then be made by connecting the points so identified on the cdf's. Figure 20 shows an example of very little movement of impulse noise level. Figure 21 shows an example of very great movement. In Figure 20 the impression is that the impulse noive level

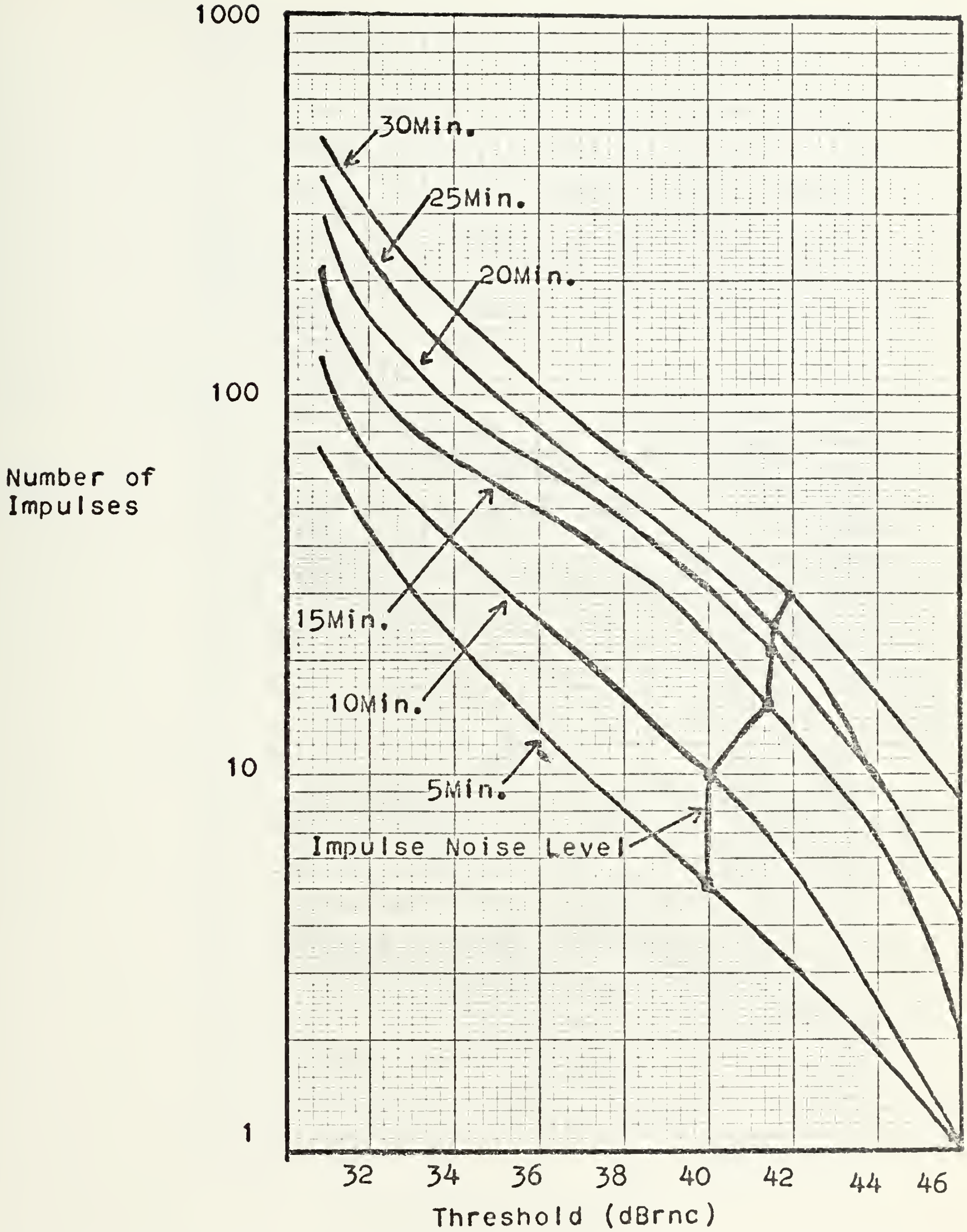


Figure 20- Cumulative Distributions of Impulse Noise Peaks In Successive Intervals From 5 to 30 Minutes.

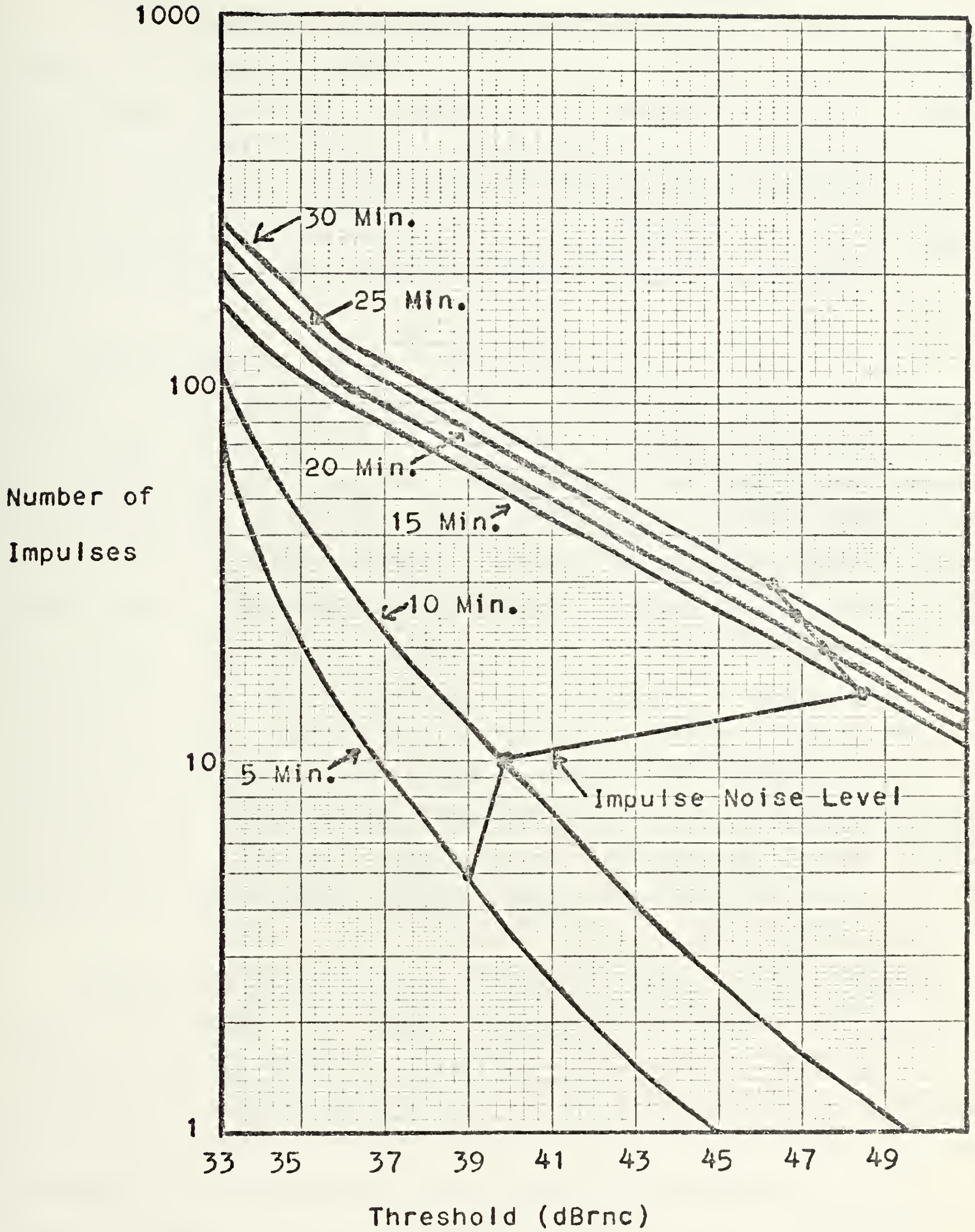


Figure 21- Cumulative Distributions of Impulse Noise Peaks In Successive Intervals From 5 to 30 Minutes.

is "settling down" by the end of the 30 minutes. Many such sets of cdf's were constructed and examined. In almost all cases, there was a qualitative feeling that the 30-minute measurement provided a reasonable estimate of impulse noise level. At least it is an average taken over 30 minutes and so is a better estimate than that provided by a shorter one. This is the only known supportive evidence that 30 minutes is an adequate measurement interval for impulse noise.

Most impulse noise originates with the actions of people. Subscribers initiating and terminating calls cause relays and switches to be activated and released giving rise to impulse noise from the associated electrical transients. Normal maintenance, installation, and repair activities, within the communication system introduce impulse noise. Since people use the communication system more during the day than at night, impulse noise exhibits large diurnal variations. Thus it makes little sense to talk about the impulse noise level on a circuit. What is obtained in a measurement made during the normal working day is an estimate of the maximum impulse noise level achieved. With this thought in mind, Figure 21 now represents a 30 minute trace of impulse noise level during its peak period.

Using the premise that a 30-minute interval provides a usable estimate of peak impulse noise activity, the next logical step is simply to use an even shorter interval. The penalty in uncertainty incurred by using shorter intervals must be examined. Using the 30-minute estimate as the yardstick and looking at the relative variance of estimated impulse noise level as a function of time, two examples of the change in the variance of estimated level, compared to the 30-minute estimate, are presented in Figures 22 and 23. Referring to Figure 23 it is noted that the variance falls from 5.0 dB to 0.1 dB as the interval increases from 5 to 25 minutes. Since the distribution of noise levels is normal

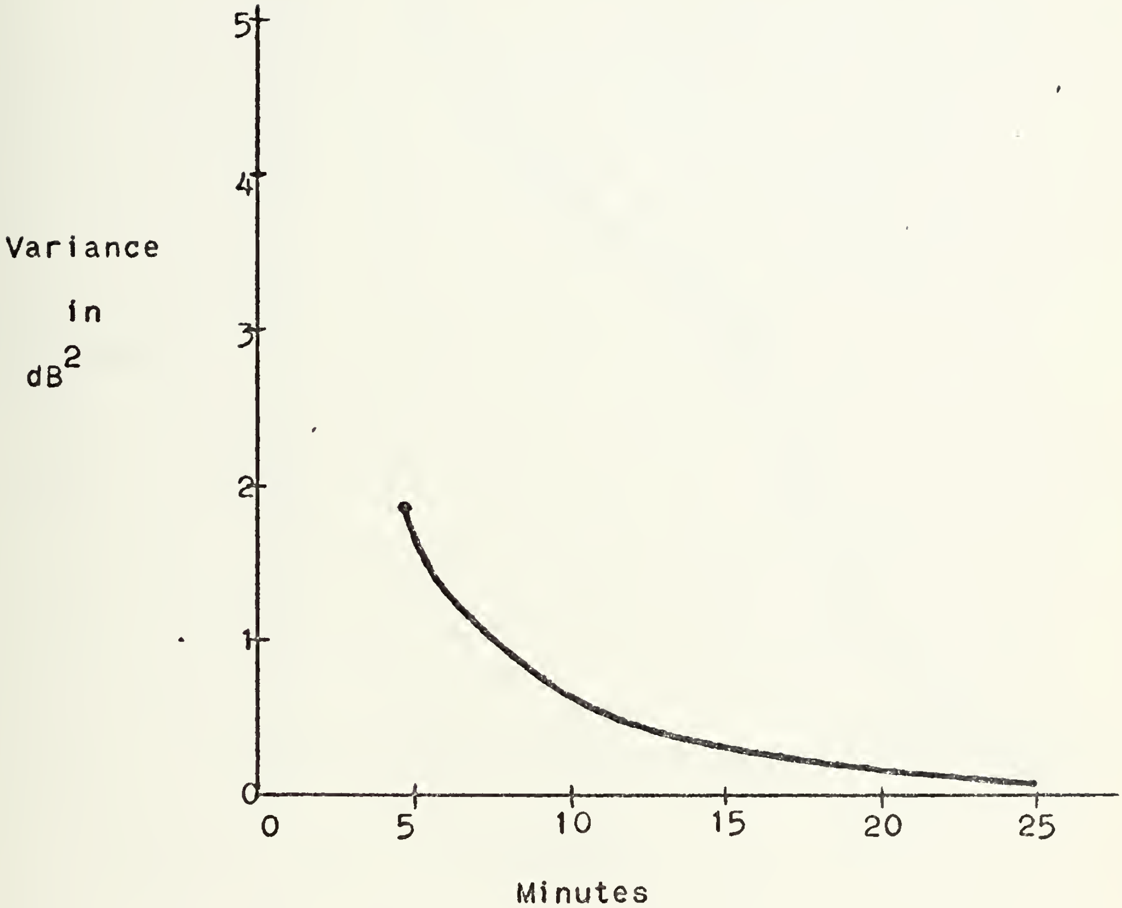


Figure 22- Variance of Estimated Impulse Noise Levels In Increments From 5 to 25 Minutes When Compared With A 30 Minute Estimate- Cable Carrier Example.

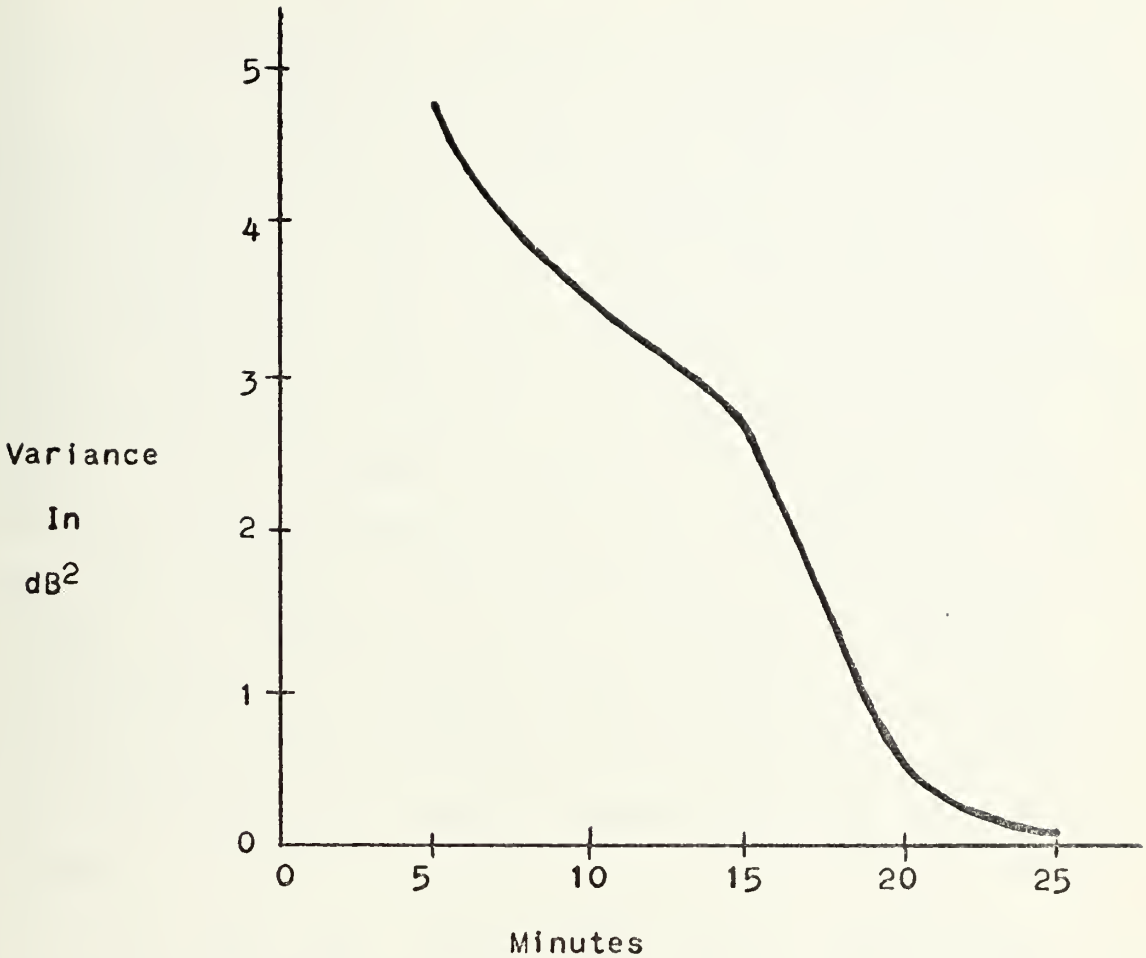


Figure 23- Variance of Estimated Impulse Noise Levels In Increments From 5 to 25 Minutes When Compared With A 30 Minute Estimate. (Radio Facility Example)

the expected error in a measurement of from 5 to 25 minutes long can be derived from the data in Figure 23. The standard deviation for 5-minute estimates is about 2.2 dB so it can be determined that 95 percent of all 5-minute measurements will be within ±3.6 dB of a 30-minute measurement. Similarly, 95 percent of the 15-measurements will be within ±2.7dB of a 30-minute measurement. Most measurements will come closer to the 30-minute estimate than the numbers just given. For example, using the data in Figure 22 it is found that 95 percent of 5-minute estimates will fall within ±2.3 dB of a 30-minute one. Impulse noise level is not simple to measure. It normally requires the use of a multithreshold counter and interpolation along a cdf of counts at some specified threshold or a maximum number of counts at some specified threshold. The requirements are always based, however, on a desire to control the maximum impulse noise level. As shown next, estimates of the 1-count per minute impulse noise level may be made solely from count distributions at an arbitrary level.

It has been stated that count distributions are log-normal and that impulse noise level distributions are normal. There are, however, simple relationships to change from count distributions. The mean of the normal impulse noise level distribution is simply the value, in dBrnC, at which the impulse noise test set recorded the count distribution. It has a count associated with it which is simply the median of the observed count distribution. The standard deviation of the impulse noise level distribution may be estimated by the expression $\sigma_{\ell} = m\,\sigma_{0}$, where m is the inverse slope of the peak amplitude distribution in dB per decade of counts (averaging 7.0; see Figure 24) and the standard deviation (σ_{0}) of the log-normal count distribution is calculated by taking the square root of the log of the ratio of average number of counts to the median count.

Thus:

$$\sigma_0 = (\log_{10} (\text{average count}/\ \text{median count}))^{1/2}$$

Where the median is not equal to zero.

In the evaluation of impulse noise level on channels, the average and standard deviation of the impulse noise level may be estimated as above. Let the impulse noise level from a set of measurements be N counts per unit time. The threshold at which the median of a number of measurements would be one count per minute can be estimated by using Figure 24 which shows the peak amplitude distribution of impulse noise average over many facilities. For example, if N=10 counts per minute at a given threshold, the one count per minute threshold would be 7 dB above the threshold used to measure 10 counts per minute.

The impulse noise level on a transmission facility may be determined by making 5-minute measurements on a number of channels on the facility and computing N as discussed above. By the use of Figure 24 the measurements may be taken at any threshold and converted to the one count per minute threshold. Knowledge of the standard deviation of the impulse noise level is also useful in engineering facilities. For example, if the standard deviation is large, more margin against the effects of impulse noise may be provided by reducing the impulse noise level below normal objectives.

While the net noise on channels in a frequency division multiplexed carrier facility will be essentially the same, in general it cannot be expected that an external transient will cause a simultaneous pulse on all channels. The noise is introduced to each channel by a process which

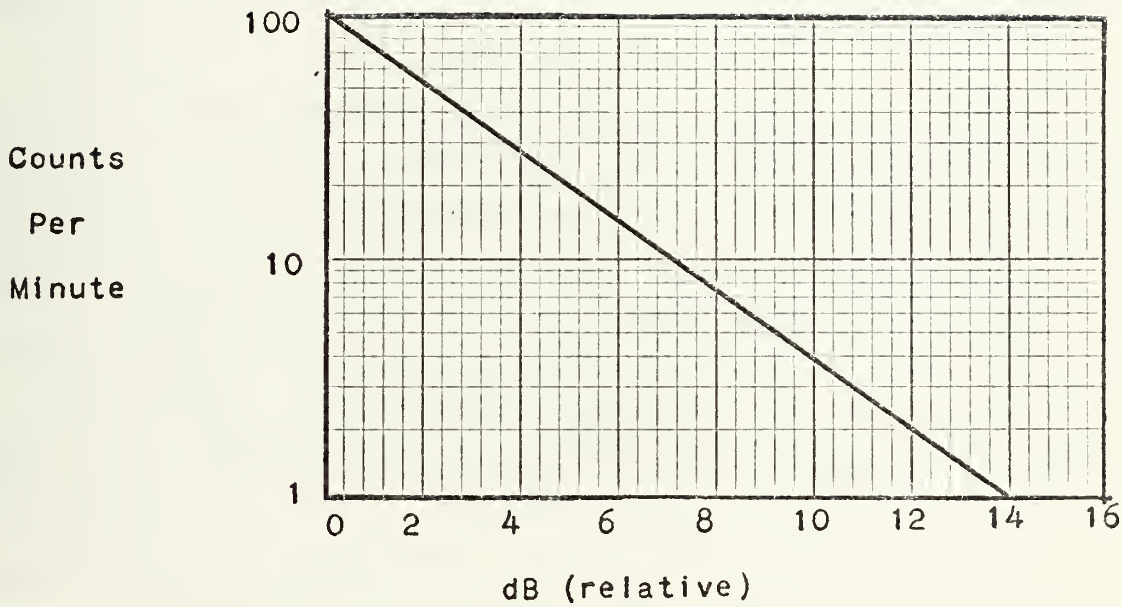


Figure 24- Average Impulse Noise Peak Amplitude Distribution.

essentially multiplies the external pulse with each channel carrier signal. In general, the carriers are not coherent, i.e., they do not all go through maxima and minima at the same time. Hence, one external influence will produce observable pulses on some channels but not on others. This same process also causes external transients of equal magnitude to be dispersed in amplitude on the carrier channel. Thus, fixed amplitude disturbers will be spread in amplitude over a range of about 20 dB on a double sideband carrier system and about 6 dB on a single sideband carrier system when measured at the output of a single carrier channel.

It is known that impulse noise frequently occurs in clusters. Due to size and cost considerations, early instruments designed specifically to count noise pulses used electromechanical counters. It was recognized that some number of noise pulses would be missed because they occurred shortly after a noise pulse in the process of being recorded by the counter. The time it takes a counter to register one count is referred to as dead time. In order to collect comparative data with different sets, the dead time is electronically controlled to be very close to 140 ms. In order to evaluate the effects of not counting all pulses, comparative tests were run using an electronic counter and an electromechanical one. Distributions of the number of pulses missed in a 30-minute period due to dead time were constructed. Figure 25 shows that in some instances as many as 120 may be missed because of the dead time. From data such as shown in Figure 25 it is possible to use an average number of missed impulses as a correction factor. It is, however, simpler to include such a factor in stated objectives for control of impulse noise rather than force the use of a correction factor after each measurement. Even though the distribution of missed noise pulses is very skew, research has not uncovered any problems with this

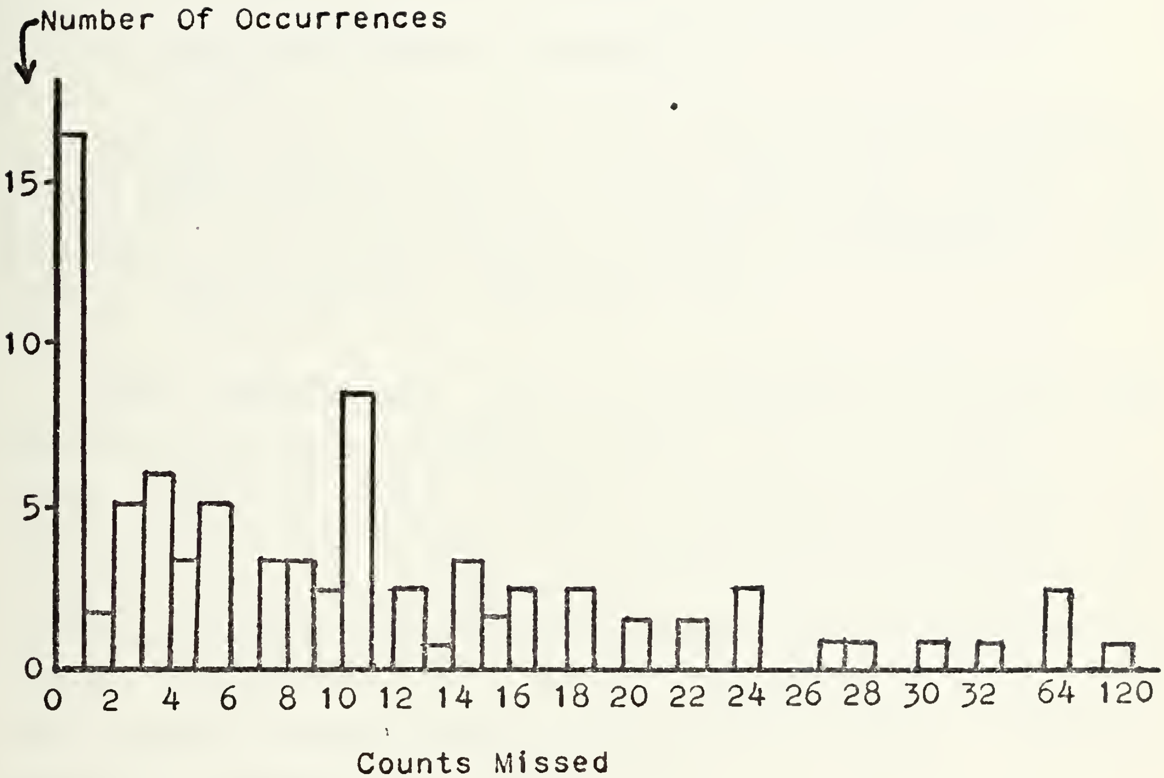


Figure 25- Number Of Impulses Missed In A 30 Minute Period By Electromechanical Counters. Threshold Always Adjusted To Produce Exactly 30 Counts In 30 Minutes On Counters.

approach. Even though the number of missed impulses with an electromechanical counter may appear gross, the change in threshold (increased sensitivity) required from the electromechanical counter to achieve a count equal to the electronic counter is only about 0.9 dB on the average, with a standard deviation of 0.76 dB. (Reference 28)

3. Single Frequency Interference

Single frequency (SF) interference refers to unwanted steady tones. Short bursts of tones which may occur from crosstalk of multifrequency signaling, for example, do not fall in this category. (Reference 22)

The requirements for SF interference is that, when measured through a C-message filter, it be at least 3 dB below C-message noise limits. A simple audio monitoring arrangement on the output of a C-message noise test set will usually detect this interference since tones exceeding the limit are easily heard if the C-message noise is within its normal range. The audio monitoring arrangement eliminates the need for sweeping the channel spectrum with a frequency selective voltmeter to detect tones.

A foolproof test for SF interference is not possible with either technique. If the SF tone source occurs before a compandored facility, chances are that it will not be detected with a C-message noise measuring set since it will be down approximately 18 dB from its power in the presence of a signal. Audio monitoring at the output of a C-notched noise test set will detect SF interference for voiceband tones not in the rejection band of the notch filter.

Single frequency interference is potentially most disturbing to systems which frequency multiplex several narrowband channels on one voice bandwidth channel. Voice

frequency telegraph equipment is a common example. Consider a system which subdivides a voice channel into 20 telegraph channels. The C-message noise in each narrowband channel will be down from that measured in the full voiceband channel by about 13 dB. SF tones, however ,do not realize the reduction due to the narrower bandwidth and may be the controlling impairment in such a situation. This fact should be taken into account when multiplexing is used.

4. Quantizing Noise

To transmit analog signals over digital lines, the analog signals are first sampled in time. One of a finite number of amplitudes is then used to represent each sample to the nearest possible amplitude. The distortion caused by the round-off error, in conjunction with sampling, is called quantizing noise or quantizing distortion.

Quantizing noise is measured in the same way as C-notched noise as previously described. Although quantizing noise is signal correlated, it is considered to have the same effect on transmission quality as an equal (in power) amount of background noise. This is somewhat conservative since tests have indicated that the impairing effect is 0 to 3 dB less than Gaussian noise. Quantizing noise can appear as other impairments when viewed through certain test sets. For example, it can appear as harmonic or intermodulation distortion. It may also appear as phase jitter. (These parameters will be discussed in detail in the section on distortion.)

The measurement of quantizing noise reflects an additional component. System imperfections on PCM systems can cause nonlinear distortion. Because of the sampling, the out-of-band energy from this nonlinear distortion is folded over and appears in-band. Thus, when a 2800 Hz

holding tone is used to measure quantizing noise ,the second and third harmonic distortion caused by system imperfections appear at 5600 Hz and 8400 Hz. These tones are folded over and appear at 2400 and 400 Hz (assuming a sampling rate of 8000 samples per second) and add to the measurement since they are not removed by the 2800 Hz notch filter at the receiver.

5. Summary

In summary the important parameters with regards to noise are:

a. Message Circuit Noise: Background noise measured through a weighting network.

b. C-Notched Noise: A measure of the background or impulse noise measured through a weighting network when a holding tone, usually 2805 Hz at -10 dBm0, is being transmitted over the system under test. The tone is blocked at the measuring set by a notch filter.

c. Impulse Noise: Indicated by the number of noise bursts exceeding a selected voltage threshold.

d. Single Frequency Interference: Spurious tones present on the channel in addition to the desired signal.

e. Quantizing Noise: Signal correlated noise generally associated with the quantizing error introduced by analog-digital and digital-analog conversions in digital transmission systems.

C. DISTORTION

Distortion exists when an output wave form is not a true reproduction of the input wave form. The imperfections and inherent non-linearities of the components of the communications system will not allow the output to be an exact reproduction of the imput, thus distortion results. Distortion may consist of irregularities in amptitude, frequency, or phase.

1. Envelope Delay Distortion

It is difficult to measure the phase characteristic of a transmission system since a phase reference is hard to establish at the receiving end of the circuit and the channel may have a varying zero frequency phase intercept . Because of this, the derivative of phase, the envelope delay ($d\phi/df$) ,has been used as a measure of the phase linearity of circuits. In practice, the true derivative cannot be measured either but can be approximated by measuring the difference in shift ($\Delta\phi$) experienced by the sidebands of a narrowband AM signal, and presenting the approximate derivative, $\Delta\phi/\Delta f$,on an instrument. The quantity Δf is twice the modulating frequency and is referred to as the aperture of the instrument. The difference between the quantity $\Delta o/\Delta f$ measured at some frequency, and that measured at some reference frequency is referred to as envelope delay distortion.

The phase characteristic of any system can be represented by a constant plus a linear term plus the Fourier expansion of the phase distortion. Thus:

$$\phi(f) = A_o + af + \sum_{i} A_i \operatorname{Sin}(2\pi \tau_i f)$$

Where $1/\tau_i = f_i$ is the period of the i^{th} phase component in Hz.

Measured envelope delay, D, is related to the true derivative of the phase by the expression:

$$D = \Delta\phi/\Delta f = ((\mathrm{Sin}\ x)/x)\ (d\phi/df)$$

Where $x = (\pi\Delta f)/f_i$.

When $f_i = \Delta f/n$ (for n = 1,2,3,....), then D=0 so that the instrument is blind to the components equal in period to submultiples of the aperture of the set. As the component periods increase (Sin x)/x approaches 1 and D approximates the derivative more closely. Thus there exists a weighting of the phase derivative by the measuring instrument which is a function of the component period and the aperture of the set. Some weighting is desirable as shown below.

Consider one component of the phase characteristic to be given by $\phi(f) = A(\mathrm{Sin}\ 2f\pi)/f_o$. If A is small, as it usually is, this results in one leading and one lagging echo of the transmitted signal which will cause intersymbol interference. The magnitude of the echoes will be given by A/2 and they will be displaced in time from the original signal by τ_o seconds ($\tau_o = 1/f_o$). The derivative of the above phase curve is $(A2\pi/f_o)\cos(2f\pi/f_o)$ in the delay domain. Note that the amplitude of the component in the delay domain is changed by the factor $2\pi/f_o$ and so echo

magnitude is not immediately evident from the delay curve. With A held constant and f_o variable, Figure 26 demonstrates the delay amplitude, $(A2\pi)/f_o$ as a function of f_o where the contours indicate equally interfering delay component peaks. The magnitude of the delay component, for echoes of equal magnitude, increases as f_o decreases. If a true phase derivative instrument could be built, small period components resulting in echoes equally as interfering as large period components would tend to completely obscure the delay plot. Thus some smoothing of small period components is desirable in order to keep the observed delay distortion irregularities in proper perspective.

Several different apertures have been used in practice but this discussion will be limited to two. One is the current Bell System standard of 166-2/3 Hz and the other is the international standard (CCITT) of 83-1/3 Hz. Figure 27 shows how each of these apertures modifies the true derivative of the phase characteristic by the factor (Sin x)/x. The factor (Sin x)/x is plotted as a function of f_o for each of the two apertures. Echoes are usually of importance only where part or all of the transmission path consists of 2-wire facilities. The plots have been truncated at f_o = 20 Hz because it is the practice to install echo suppressors on 2-wire connections when the round trip delay exceeds 45 milliseconds, and f_o is the reciprocal of the round trip delay or echo time. Thus components finer than 20 Hz should be of no interest.

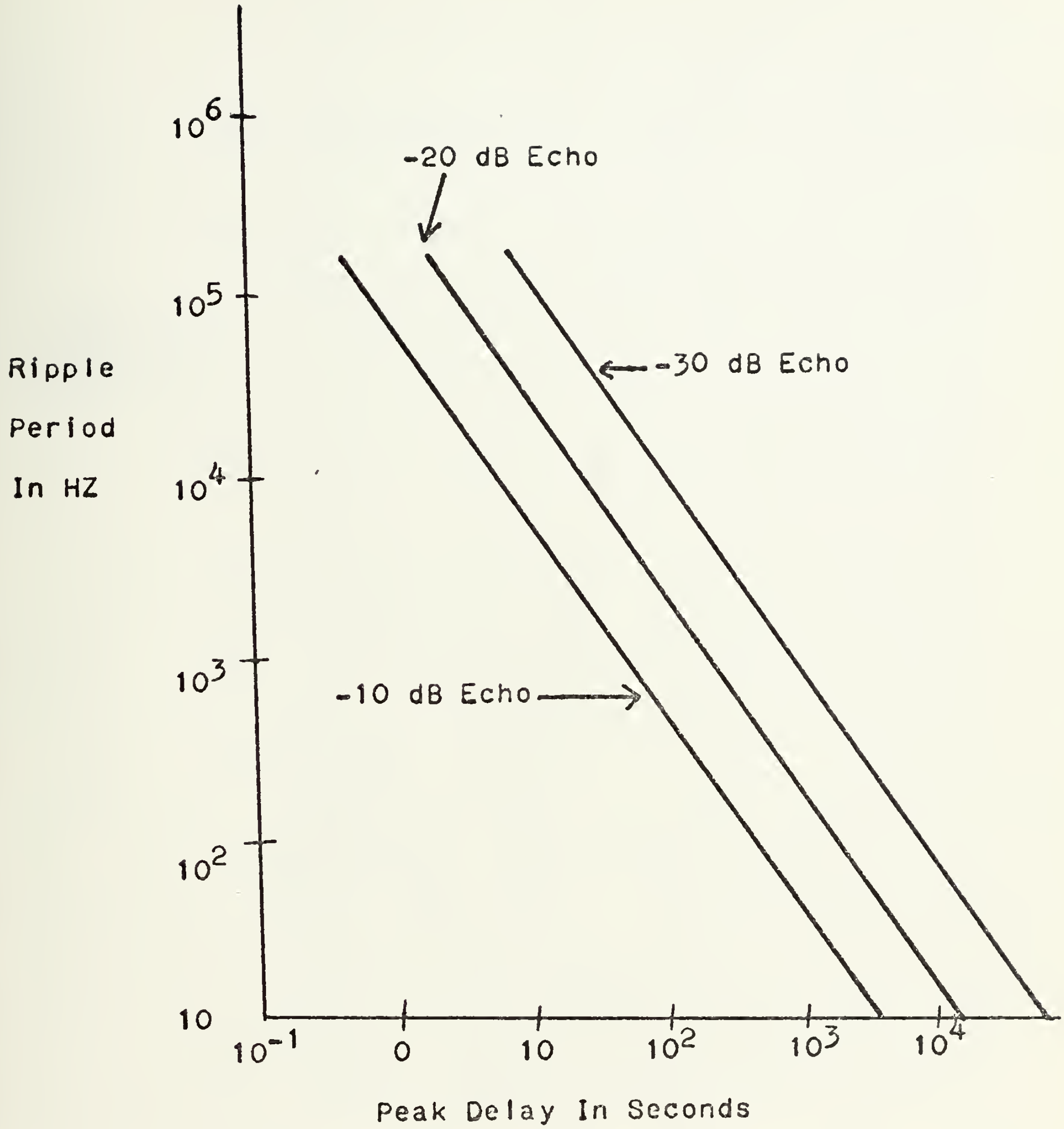


Figure 26- Peak Delay For Single Sinusoidal Ripple With Constant Echo Magnitude.

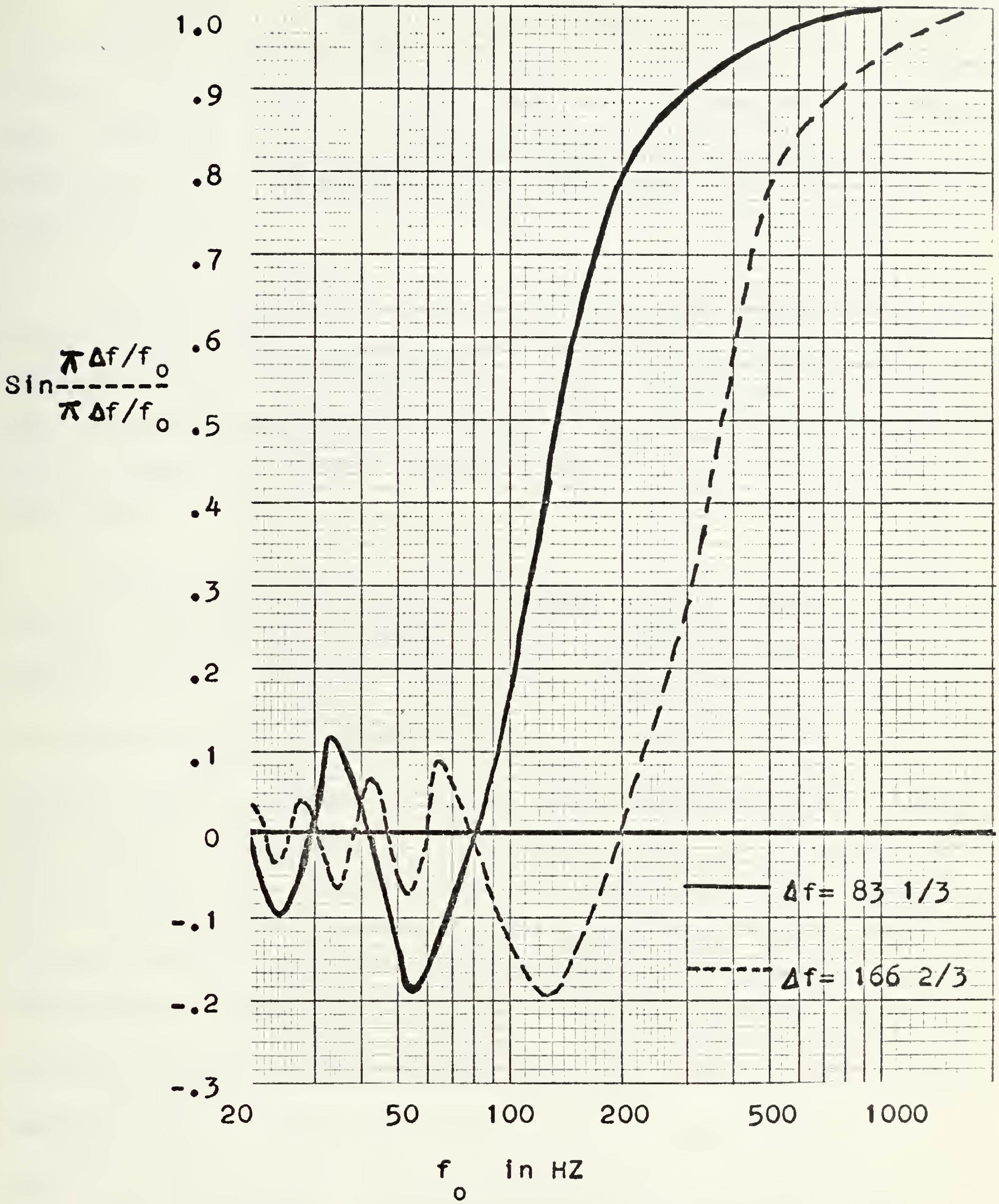


Figure 27- Response Of Two Envelope Delay Test Sets To Ripples In The Phase Characteristic Of A Transmission Line.

One exception to this rule will occur if echo suppressors are disabled to permit simultaneous operation in both directions over a single 2-wire connection. In practice, echoes with a delay time of about 45 milliseconds or greater are usually not a problem because of the extra loss they encounter in traversing the relative long echo path. (Reference 22)

From Figure 27 it can be seen that the Bell System sets have eight blind spots (zero crossings) and come within 90 percent of the true delay for ripples at 500 Hz and higher. The international standard aperture has four blind spots and comes within 90 percent of the true delay at about 300 Hz and above.

The magnitude of the intersymbol interference, due to a single sinusoidal component of the phase characterstic, is given by $A_i/2$. The peak possible intersymbol interference is given by $\sum_i |A_i|$, where the A_i are the coefficients of the Fourier series expansion of the phase characteristic across the band of interest. In theory, this could occur for a given phase characteristic for a particular signal sequence such that, at one instant of time, all the echoes generated by the preceding signals added in phase. Because of the importance of the quantities A_i is is of interest to examine possible errors in the extraction of the A_i from delay measurements. This can be done by examining how the A_i are modified by a perfect envelope delay measuring set and by a realizable envelope delay measuring set. (Reference 29)

A typical example of the usual manner of specifying delay requirements is shown in the following table :

TABLE II

Permissible Envelope Delay Distortion Limits	Within Frequency From	To
500 Microseconds	1150	2300
900 Microseconds	1000	2500
1750 Microseconds	800	2700

In addition, listener echoes should not exceed a power of -12 to -18 dB with respect to the received signal. This means for example, that a distortion consisting of a single isolated echo should be no larger than one-fourth of the main signal. Such echoes are proportional to the peak-to-phase deviations on the channel. However, the peak-to-peak envelope delay distortion is proportional to the product of the amplitude of an echo and the time displacement between the echo and the main received signal $1/f_i$ (Peak-to-peak envelope delay is proportional to A/f_i). Because of this, echoes having long delay paths tend to cause very large envelope delay distortion even though they may cause very little difficulty to data transmission. This effect is shown by curve I of Figure 28. As the curve shows, 12 dB listener echoes that are delayed more than a few milliseconds cause peak-to-peak envelope delay excursions that greatly exceed the 500-microsecond requirement of Table II. If their delay were on the order of 10 to 20 milliseconds, such echoes could completely obscure a plot of envelope delay distortion even though they would not be significantly interfering. Fortunately,

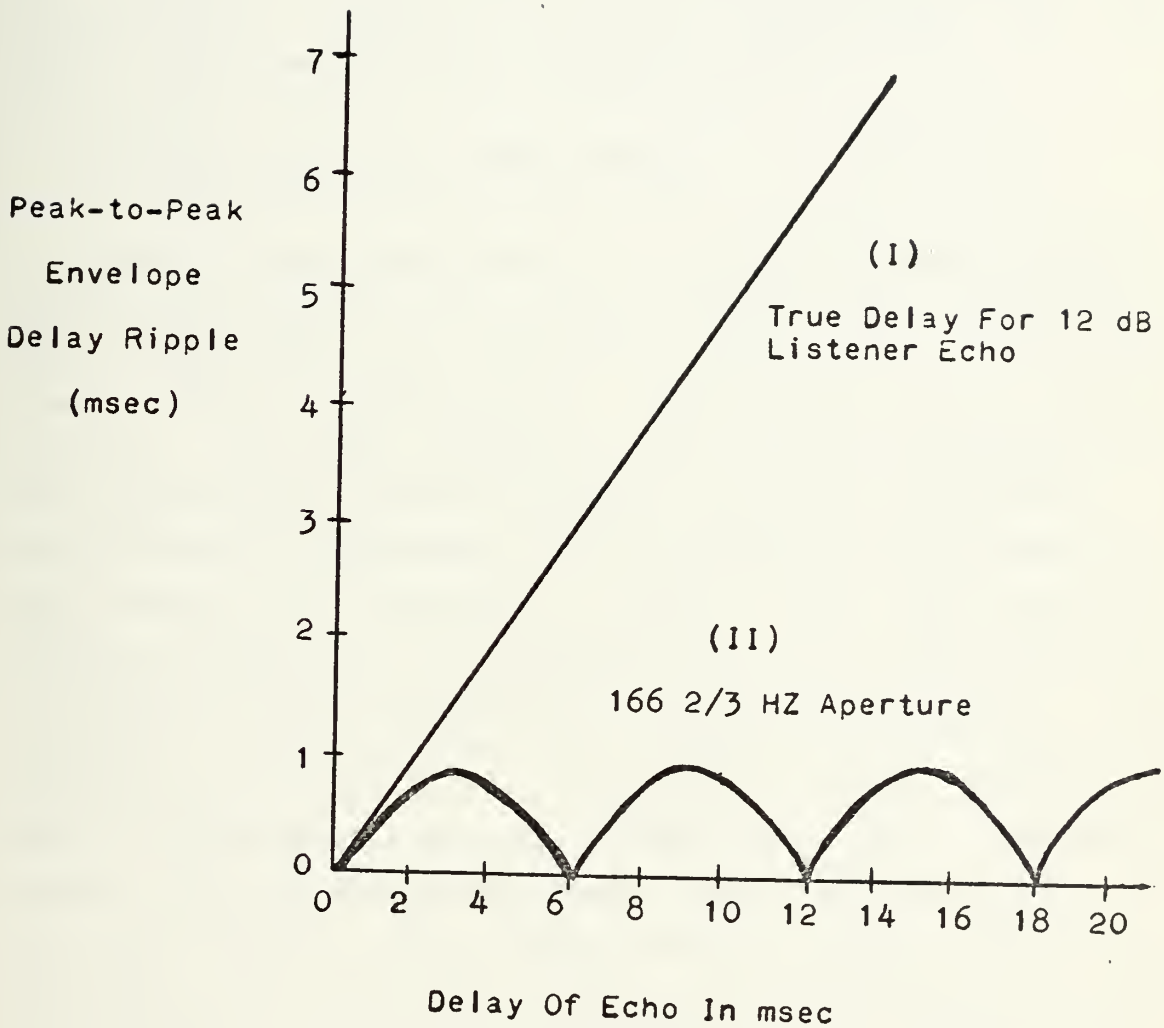


Figure 28- Response Of Envelope Delay Test Set To Echos As Compared To True Delay.

the aperture error determined by the modulating frequency of a delay measuring set will tend to offset this effect. Curve II of Figure 28 shows the response of the delay set having an aperture of 166-2/3 Hz to a 12 dB listener echo. The response of the delay set, while having peaks of 1000 microseconds, remains in the neighborhood of 500 microseconds peak-to-peak. (Reference 30)

2. Peak-To-Average Ratio (PAR)

Due to the difficulties in evaluating the resultant intersymbol interference from an envelope delay measurement, it is difficult to establish requirements in the delay domain which satisfactorily characterize the channel quality. To help overcome this problem, a technique has been developed to measure channel dispersion (spreading in time of signal amplitude) due to transmission imperfections. The concept is to generate a pulse train with spectral content shaped to be represenative of a data modulated voice-band signal with spectral components chosen at the generator to give rise to a high Peak-To-Average Ratio (signal peak to full wave rectified average, abbreviated PAR) of the signal. As such a signal traverses a dispersive medium, the PAR will deteriorate. Then by measuring the PAR at the receive end, a simple measure of the dispersion is obtained. If the prime source of dispersion is phase nonlinearity, as it is in telecommunication channels, then a quick measure of this impairment is possible.

The test set used to measure PAR also responds to other channel impairments but to a lesser extent than phase nonlinearity. Frequency response and nonlinearities are the second and third most important degradations followed by C-notched noise, and incidental FM.

The PAR test signal has a spectrum centered at about 1650 Hz and has 12 dB points at about 1000 and 2400 Hz. The detector responds approximately according to the relation:

$$PAR = 100\ ((2E(pk)/\ E(FWA))-1)$$

Where E(pk) is the normalized peak and E(FWA) is the normalized full wave rectified average of the envelope.

In general, PAR ratings of about 50 or higher indicate that intersymbol interference for medium speed (2400 bits per second) data transmission will be acceptable.

The PAR ratings of a system may be calculated from the gain and phase or gain and delay characteristics of the channel if it is desired to know the rating percisely. The PAR test is used as a quick method of evaluating telecommunications channels for nonvoice transmission. It provides a rapid means of sorting between channels having acceptable and unacceptable phase characteristics. The accuracy of the meter is typically ± 1 PAR point and deviations in excess of 8 points from a calculated expected value, or ± 4 PAR points from an initially measured value, provides sufficient reason to supect that some characteristic has changed significantly. (References 31,32,33 and 34)

3. Nonlinear Distortion

Nonlinear distortion can be broadly defined as the generation of signal components from the transmitted signal that add to the transmitted signal, usually in an undesired manner. The main sources are electronic devices and other components comprising voiceband channels. The nonlinear distortion of concern here should not be confused with the

intermodulation noise caused by nonlinearities in the terminal equipment and line amplifiers of a frequency division multiplex system. Although these nonlinearities can contribute to the nonlinear distortion at voice frequencies, their contribution is usually negligible.

Nonlinear distortion is commonly measured and identified by the effect it has on certain signals. For example, if the signal is a single tone of frequency A, the nonlinear distortion appears as harmonics of the input. Thus it appears as tones at 2A, 3A and so on. Since most of the distortion usually occurs at the second and third harmonic, it is often measured by the power of each of these harmonics, and is called second and third harmonic distortion. If the amount of nonlinear distortion is measured by the power sum of all the harmonics, the result is called total harmonic distortion. These distortion powers are not meaningful unless the power of the wanted signal (the fundamental) is known, so measurements are usually referred to the power of the fundamental and termed second, third, or total harmonic distortion.

For a multitone input signal, the nonlinear distortion is termed intermodulation distortion and appears as tones at frequencies which are linear combinations of the the input frequencies. As an example, if the input consists of three tones at frequencies A, B and C, the nonlinear distortion appears at frequencies $k_1A \pm k_2B \pm k_3C$ where k_1, k_2, and k_3 are nonnegative integers. The distortion is called second order distortion for those nonnegative values of k_1, k_2 and k_3 such that $k_1 + k_2 + k_3 = 2$ and third order distortion for those values such that $k_1 + k_2 + k_3 = 3$. The distortion at

a particular frequency is called a second or third order product. The second order product for $k_1 = k_2 = 1$ and $k_3 = 0$ and is called an A+B or (or A-B) product. The third order product for $k_1 = 2$, $k_2 = 1$ and $k_3 = 0$ is 2A-B (or 2A+B). The amount of nonlinear distortion is measured by the magnitudes of various products with respect to the received fundamental. (Reference 22)

A nonlinear channel might be modeled with a third degree polynomial:

$$y = a_1 x + a_2 x^2 + a_3 x^3 \qquad (1)$$

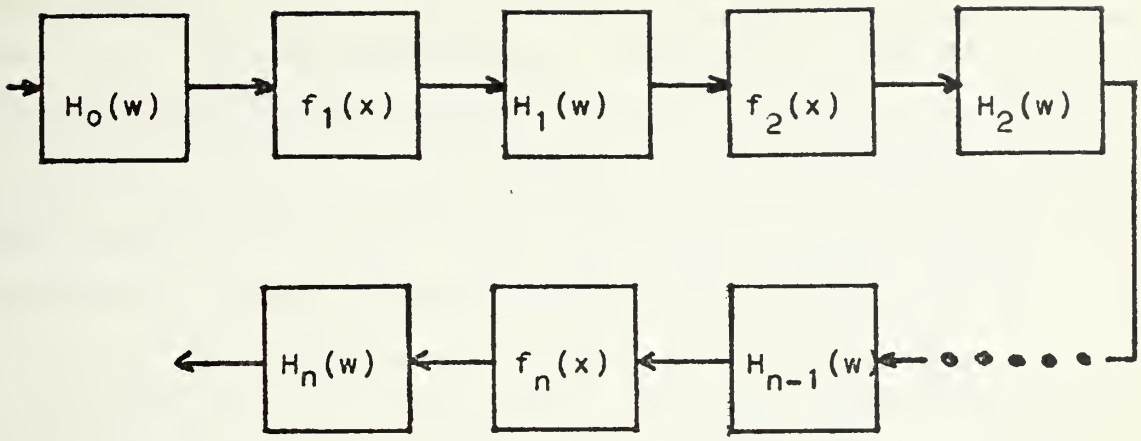
The quadratic distortion term, $a_2 x^2$, generates second order distortion and the cubic distortion term, $a_3 x^3$, generates third order distortion. If this is an accurate model, the harmonic and intermodulation distortion measurements are really equivalent for either one can be used to compute a_2^2 and a_3^3 and the measurements can be related to one another. However, because Equation (1) is only an approximation, the different ways of measuring nonlinear distortion do not always yield the same a_2^2 and a_3^3. To select the best measurement, it is necessary to understand how telecommunication channel nonlinearities differ from Equation (1). The three main differences are;

frequency dependency, time variability, and the presence of nonlinear terms above the third degree. The first two are most important and are discussed first. The third is discussed at the end of this section.

A communications channel may consist of several sources of nonlinearity alternated with linear networks as shown in Figure 29A. A simplified model for estimating transmission performance is shown in Figure 29B. This model has the same linear characteristics as the actual channel but it has only one source of nonlinearity. This nonlinearity is chosen so that the system in Figure 29B produces the same ratios of signal-to-second and -third order nonlinear distortion for a complex signal as does the actual channel modeled in Figure 29A. The nonlinear distortion measurement is intended to determine these ratios.

The amount of second and third order distortion at the output of the model in Figure 29A depends on how the distortion from the several sources of nonlinearity "adds." This addition is illustrated below using two nonlinearities, each described by $x + a_2 x^2 + a_3 x^3$ with a linear network between them.

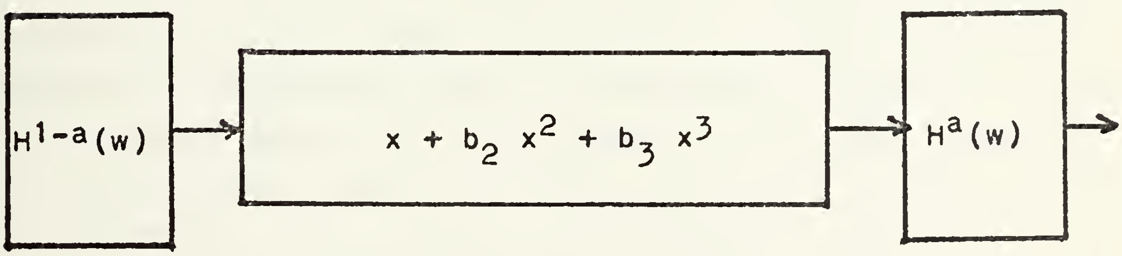
The phase angle of the linear network has the largest effect on addition of the nonlinear distortion. Envelope delay distortion is usually known about the phase angle and this is known only above some radian frequency w_o. Since the phase angle is obtained by integrating over the envelope delay distortion curve, the contribution below w_o is not known. This is left as a parameter and denoted K. On a single sideband (SSB) system K can be a function of time.



$$f_k(x) = x + a_{2k}\ x^2 + a_{3k}\ x^3$$

$H_k(w)$ = System Function Of Linear System

A. Nonlinear System



$$H(w) = {}_{k=0}^{\;n}\ H_k(w)\ , \qquad 0 \quad a \quad 1$$

B. Simplified Model

Figure 29- Two Models Of A Transmission Line With Nonlinear Distortion.

If the system introduces a frequency shift of F Hz, K varies over the range 0 to 2 and does so F times a second. K can also vary in a random fashion. Theoretical results on the response of nonlinear systems to Gaussian imputs can be used to compute the combined effect of delay distortion and K. Although data signals are not Gaussian, the addition is influenced more by the fact that signal energy is distributed across the band than by the type of signal.

Results are first presented for third order distortion. The linear network is assumed to introduce delay distortion equal to that caused by one pair of filters commonly used on SSB systems. This amount of delay distortion is shown by the curve labeled "I" in Figure 30. The in-band third order distortion is plotted along the ordinate in Figure 31 in dB with respect to the distortion generated by only one of the nonlinearities. Thus, 6 dB represents voltage addition and 3 dB power addition. K is plotted in degrees on the abscissa. A Gaussian imput having a raised cosine spectrum from 700 Hz to 2600 Hz is assumed and in-band frequencies are taken to be 700 to 2600 Hz. However, results are very insensitive to spectrum shape and bandwidth. Figure 31 also shows the level of the third harmonic for two different input frequencies and the level of a 2B -A product. The third harmonics show that the nonlinear distortion is frequency dependent since different results are obtained for different input frequencies. Also if K varies with time, the level of the third harmonic varies with time. The 2B -A product correlates best with third order distortion.

Several conditions are of interest for second order distortion but the following illustrate the important ones. One is for envelope delay distortion equal to that of the pair of filters described above. Since these occur on a SSB system, K is a function of time due to the frequency shift

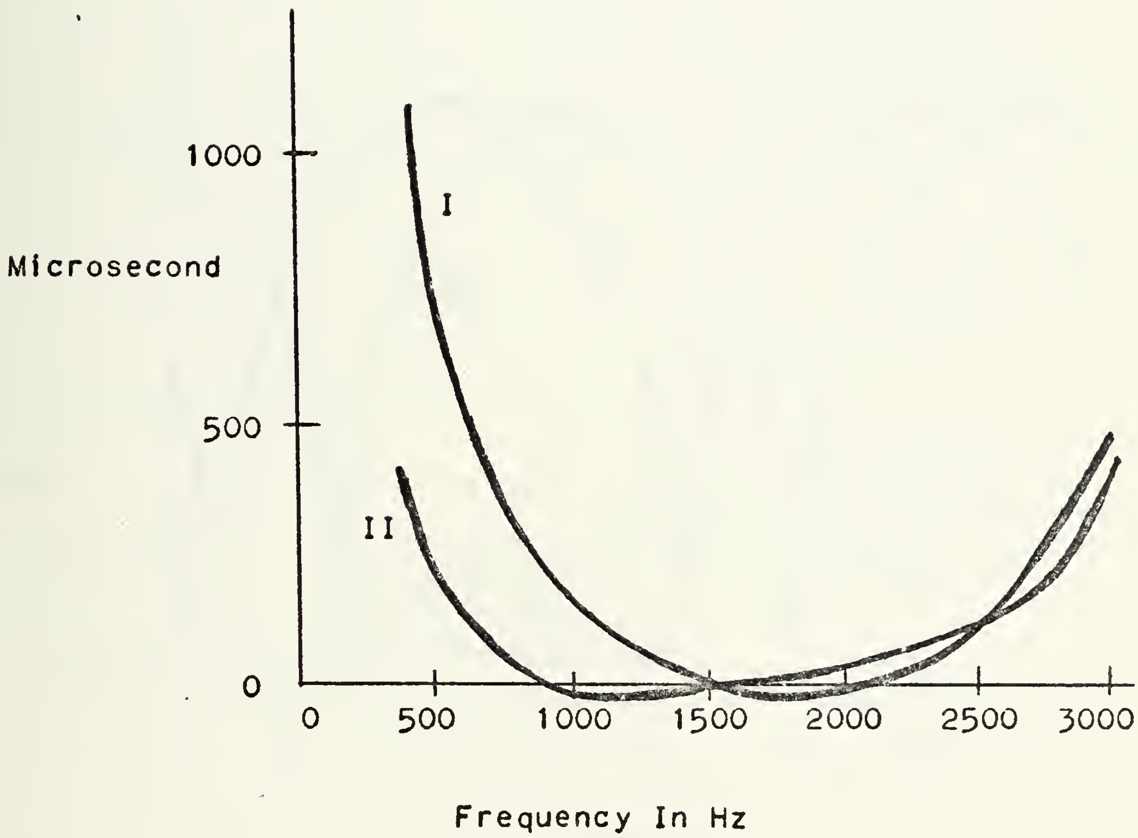


Figure 30- Envelope Delay Distortion As Measured On Two Different Transmission Lines.

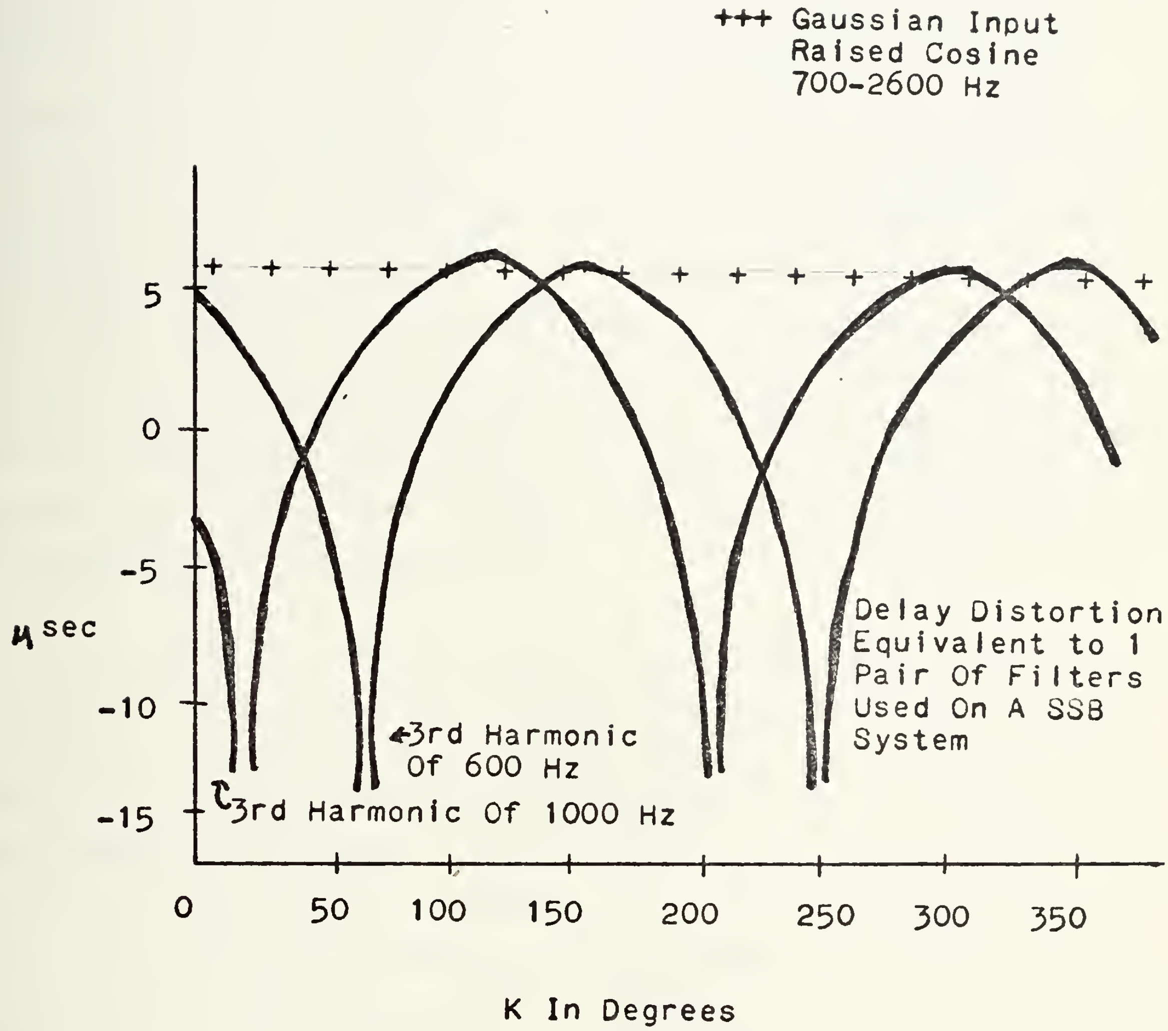


Figure 31- Some Third Order Products Compared To The Third Order Distortion For A Gaussian Input.

mentioned above. The other is for envelope delay distortion equal to that caused by another commonly used short haul carrier system. This amount of delay distortion is shown by the curve labeled "II" in Figure 30. In this case the envelope delay distortion is less and K does not change with time. Figure 32 presents results for the latter case for two different input spectra. Notice that a null exists at some values of K and that differences between the two spectra are quite large in the null. A measurement which agrees with one spectrum will not agree with the other. However, it is important to recognize that, for physical systems, differences in this region are not serious because for these values of K the second order nonlinear distortion generated at the input of the linear network tends to cancel that generated at the output. Thus the differences are large only when the amount of nonlinear distortion is insignificant. (Reference 35)

Results in Figure 33 are for envelope delay distortion caused by the pair of filters used with SSB systems. Spectrum bandwidth has a larger effect than in Figure 32 , but a null still exists at some values of K. Transmission performance in this case is a function of time since K is the function of time. However, if K is time variable there is no reason to suspect that it would spend more time at some values than at others. Thus, if the power were measured by averaging for a long enough time we would find that the second order distortion adds on a power basis. This value is labeled Equivalent Performance in Figure 33 and is used as a measure of quality for this circuit. This is the value that the measurement should indicate when the second order distortion is time variable.

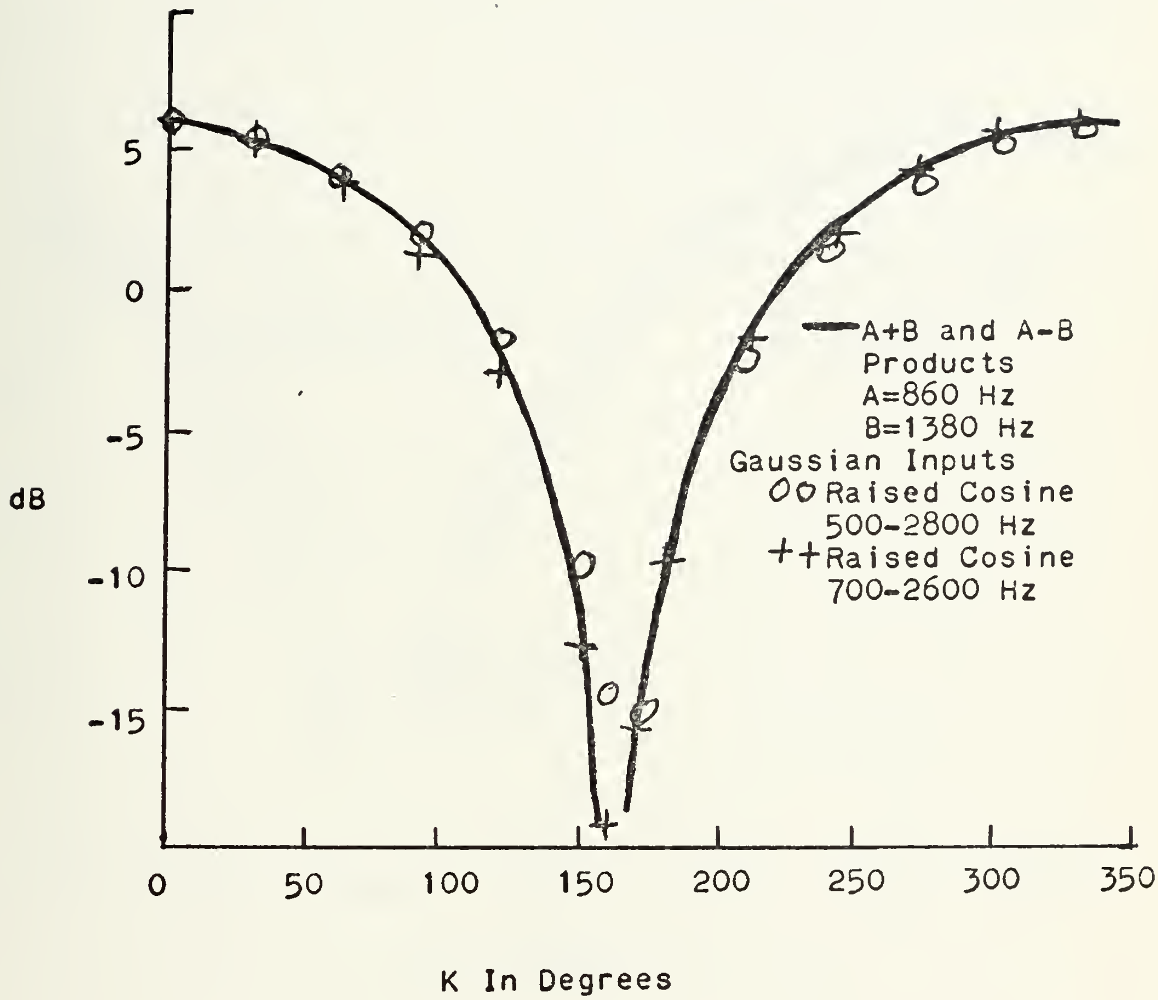


Figure 32- Power Sum Of A+B and A-B Products Compared To Second Order Distortion For Gaussian Inputs.

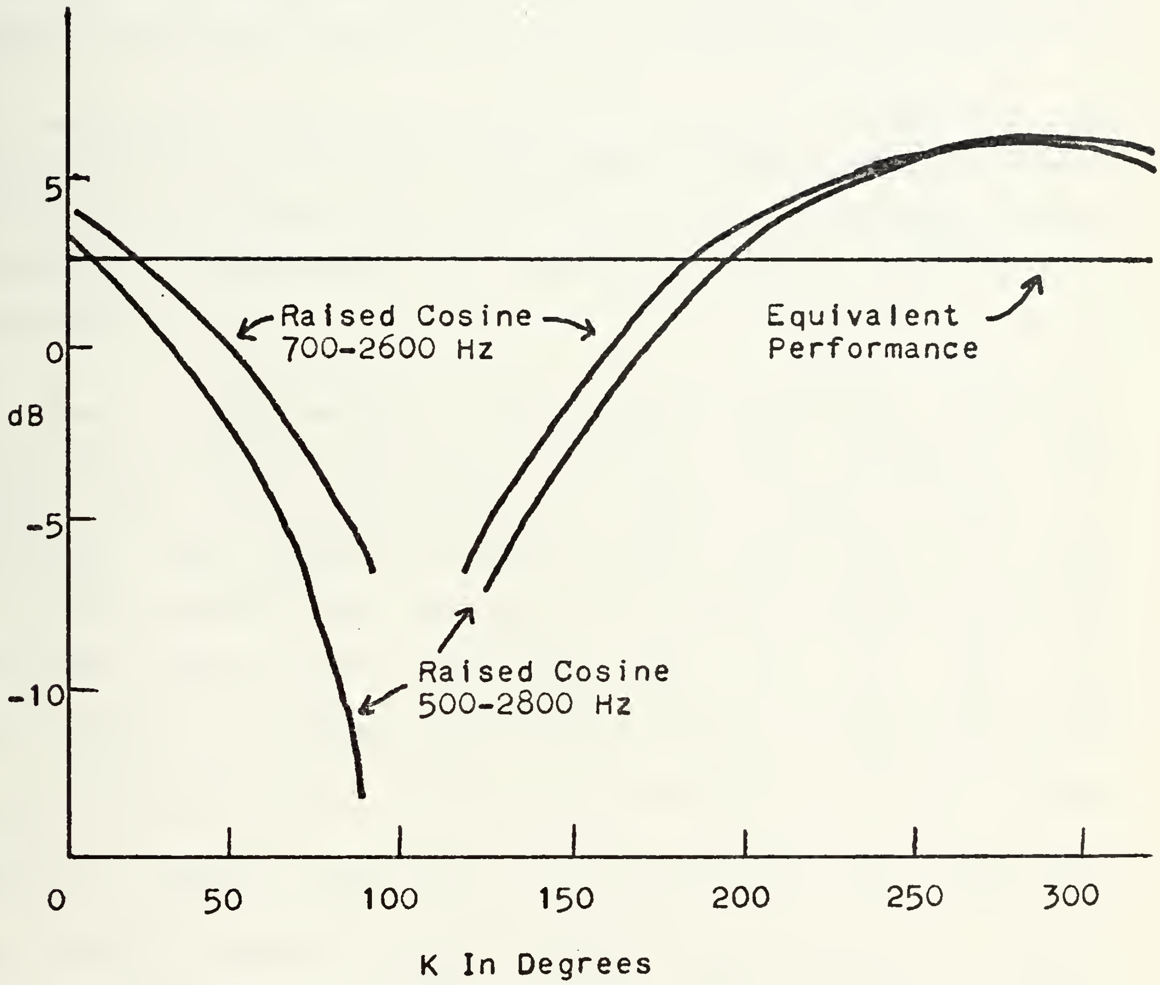


Figure 33- Second Order Distortion For Gaussian Inputs.

The models presented have demonstrated time variable and frequency dependent nonlinear distortion. It is possible that the sources of nonlinearity themselves could be frequency dependent, however, it is felt that any such effects are small over the range of frequencies of interest.

Nonlinear distortion test sets recently developed use two pairs of tones as the fundamental signal. This is known as the 4-tone method. For this test, the four equal level tones are transmitted at a composite signal power level of -13 dBm0.

The 4-tone method uses two pairs of tones. One pair consists of tones at 856 and 863 Hz (a 7 Hz spacing). The second pair uses frequencies of 1374 and 1385 Hz (an 11 Hz spacing). The frequency spacing within each pair of tones is not critical but should be different for each pair. Labeling these four tones A_1, A_2, B_1, and B_2, the second-order products (A+B) fall at A_1+B_1, A_1+B_2, A_2+B_1 and A_2+B_2. If the spacing between A_1 and A_2 is the same as that between B_1 and B_2 then $A_1+B_2 = A_2+B_1$ and these two components will add on a voltage basis and give an erroneous reading. The third order produces (2B-A) fall at $2B_1-A_1$, $2B_1-A_2$, $2B_2-A_1$, $2B_2-A_2$, $B_1+B_2-A_1$, and $B_1+B_2-A_2$. The receiver uses 50 Hz wide filters to select the A+B, B-A, and 2B-A products. R_2 is the ratio of received composite fundamentals to the power average of the A+B and B-A products. R_3 is the ratio of received composite

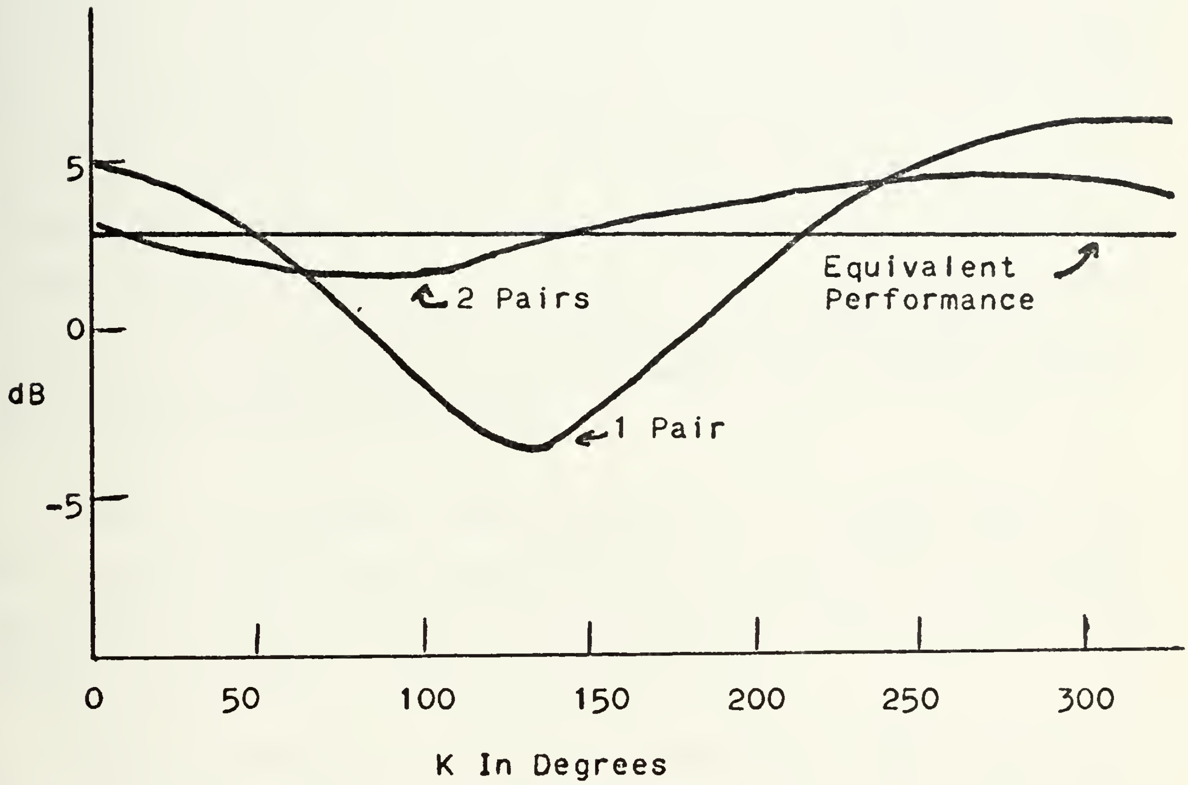


Figure 34- Power Sum Of A-B And A+B Products.

(A=860 Hz and B=1380 Hz)

fundamentals to the 2B-A products. Two second order products are used to reduce time variability The power sum of these two products are plotted in Figure 34 as a function of K for the amounts of delay distortion indicated on the figure. The measurement is not independent of K even though the depth of the null has been reduced. Thus, it is still necassary to average this reading for about 30 to 60 seconds. A slight risk exists that K will not change appreciably over 30 to 60 seconds and that the measurement will be in the null. The error could be up to 7 dB. This maximum error can be reduced by using a more complicated measurement scheme. However, more complicated schemes only reduce the maximum error by a dB or two. If the risk of being in the null is too great, after averaging 30 to 60 seconds, the measurement can be made twice, 5 minutes apart, and the largest reading used.

For the condition used in Figure 32 in which case K is not variable but the amount of delay distortion is smaller, the power sum of the A+B and B-A products tracks the second order distortion for the Gaussian inputs.

The following considerations influence the choice of frequencies A and B. As seen from Figure 31, third order distortion adds on nearly a voltage basis even with envelope delay distortion present; thus, A and B must be chosen so that delay distortion has a negligible effect on the 2B-A product. For this to happen A, B, and 2B-A should be close together and near 1700 Hz so that they appear in a relatively flat part of the envelope delay distortion curve. The A+B product is chosen less than 2300 Hz and the B-A product greater than 500 Hz to keep the effects of channel roll-off small. As B-A increases above 500 Hz, the depth of the null in Figure 34 (the maximum error) increases. Thus, B-A is kept close to 500 Hz. Finally, since phase jitter components might occur within 300 Hz of any signal

component, the nonlinear distortion products must be at least that far removed from a signal component.

The measurement must be able to identify the nonlinear distortion accurately in the presence of background noise and quantizing noise. The problem is especially severe for measuring third order distortion on part of a built-up connection because third order distortion adds approximately on a voltage basis and noise on a power basis.

Quantizing noise on a PCM system poses a special problem because when a PCM system is excited with tones, the noise spectrum is not flat and continuous but discrete. Some of these discrete components add or beat with the nonlinear distortion product being measured causing an inaccurate and time variable reading. Third order distortion is measured through a narrowband, 50 Hz filter centered at 2B-A and second order distortion through narrowband, 50 Hz filters centered at B-A and A+B. These values are then referred to the received signal level.

To protect against false measurement because of high noise or an interfering tone in one of the measurement slots the following test is made. Disable the pair of tones at B and increase the others 3 dB. This loads the system properly for a noise measurement. The part of the measurement due to noise alone can now be determined. Figure 35 is then used to determine the correction factor which must be subtracted from the distortion measurement (or added to the signal-to-distortion measurement).

It is usually thought that harmonic distortion measurements follow what is called the "power series" law, i.e., the second and third harmonic increase 2 and 3 dB respectively per dB increase of the input power. This is

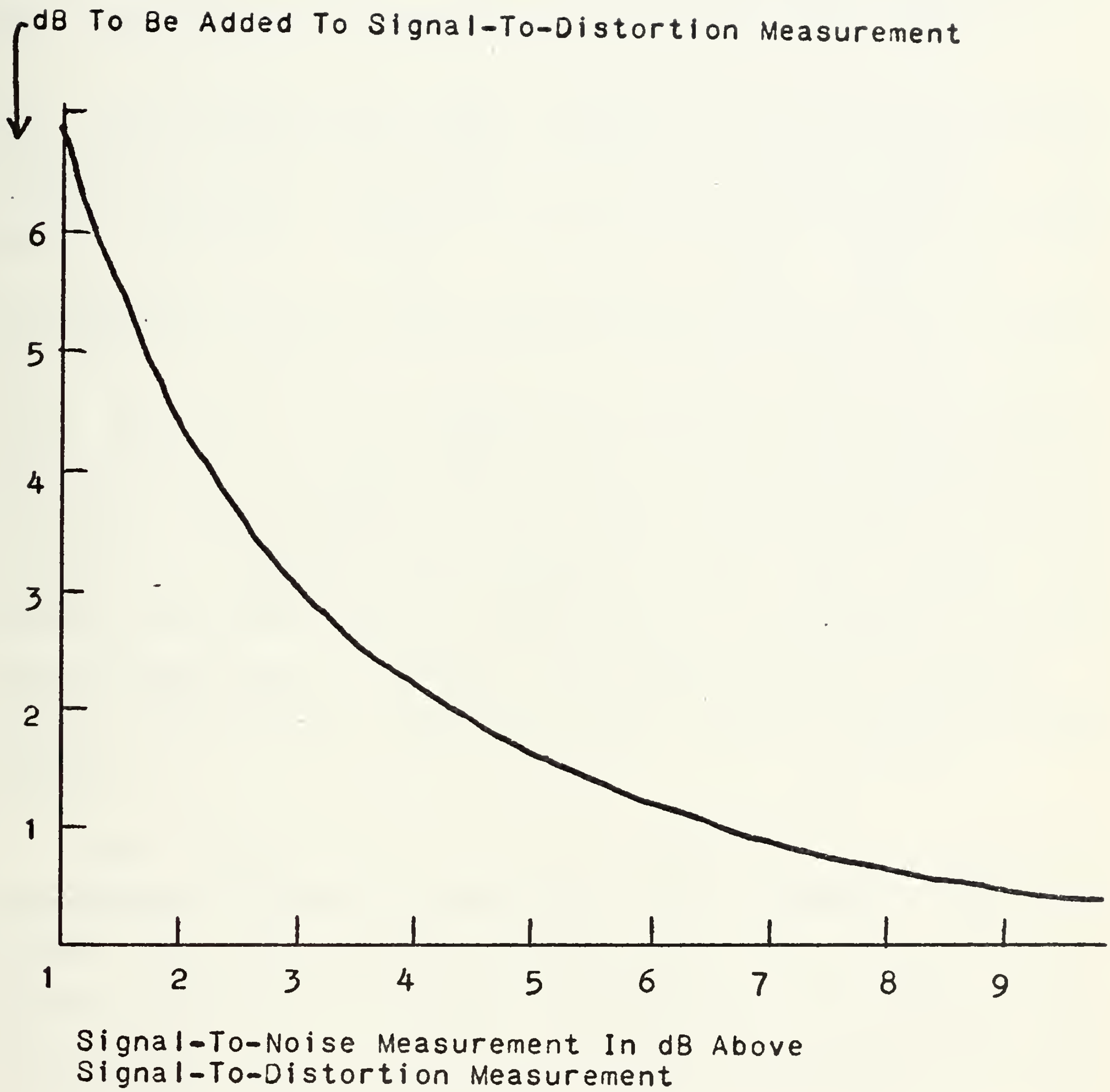


Figure 35- Correction For Noise

based on the third degree polynomial model for nonlinearities. Measurements on communications channels may not follow this law for two reasons: the presence of significant terms above third degree and the dependence of the magnitude of the nonlinearity on the input power.

An example of the first reason is a full wave rectifier nonlinearity that can occur on a PCM carrier system. The second harmonic produced by a full wave rectifier changes only 1 dB per dB change in input level. However, this nonlinearity can be well approximated with a_2x^2 if a minimum mean-squared error approximation is used. The approximation then depends on the amplitude distribution of the signal. Practically, this means that to get a good approximation the amplitude distribution of the test signal should be similar to that of a typical data signal or that they should span roughly the same range of amplitudes. For example, a 2-tone test signal should be about 2 dB above the data signal. A 4-tone test signal should have the same rms power as a data signal.

The magnitude of the nonlinearity on syllabic compandored systems depends on the compandor operating point. To produce the proper operating point, the test signal should have the same rms power as the data signal. Because of this, second and third order distortion might change only slightly more than 1 dB per dB change in input.

The signs of b_2 and b_3 in Figure 29B are not determined by the measurement. The sign of b_2 is just as likely to be positive as negative and in either case the effect on transmission performance will be the same.

However, negative values of b_3 result in a compression of signal peaks and can be more interfering to data transmission than positive values. Although it is possible that a nonlinearity may be enhancing for certain input powers, it appears that most nonlinearities encountered are compressing. (Reference 36)

4. Incidental Modulation

Incidental modulation is defined as any unwanted AM, PM, or FM imposed on the information carrying voiceband signal by a disturbing source other than itself. This is a signal correlated interference because it uses the signal as a carrier, but it is a parasitic rather than a reflexive interference like nonlinear distortion. Spurious AM imposed by electronic components with faulty power supplies and extraneous FM introduced via unstable carrier frequency sources are examples of incidental modulation. Jittering clock pulses in digital carrier systems may also contribute small amounts of phase modulation to a voiceband signal, but this is at most a second order effect.

To observe the important properties of this type of interference on frequency division multiplex systems, a single-frequency sinusoid in the center of the voiceband is transmitted. Let this signal be represented by:

$$V_t = A_o \, \mathrm{Cos}(2\pi f_o t)$$

where $1000 \leq f_o \leq 2000$ Hz

The received waveform of this tone in the absence of nonlinear distortion products can be simply expressed as follows:

$$V_r = A_o G(f_o)(1+m(t)) \operatorname{Cos}(2f_o\pi t+\phi(f_o)) + \theta(t) + n(t) \qquad (2)$$

where $G(f_o)$ and $\phi(f_o)$ are the channel amplitude and phase characteristics respectively at frequency f_o,

n(t) is the total uncorrelated interference,

m(t) in incidental AM, and

$\theta(t)$ is incidental PM

Incidental PM can be expressed in general as the following summation:

$$\theta(t) = \theta_o + (2 \quad f(t)) + \quad \theta_j \operatorname{Cos}(2 \; f_j t + \phi_j) \qquad (3)$$

θ_o, when not equal to n2 , (n=0,1,2,...) is phase intercept distortion. It contributes identical phase shift to all frequencies present in the voiceband signal. It, thereforth appears as the zero intercept on a graph of phase versus frequency for carrier derived channels. It may appear time variable due to contributions from the second and third terms.

Δf is steady-state frequency shift in Hz. It also contributes equally to all frequencies of the voiceband signal. It has the effect of shifting the signal spectrum.

θ_j, f_j, ϕ_j define N-independent sinusoidal components of the phase modulation. Note that the summation does not imply a Fourier expansion of $\theta(t)$, because the f_j are not necessarily harmonically related. Since the peak deviation in radians is usually small(ϕ_j <0.2), this constitutes low index PM with a single set of sidebands ($f_o + f_j$) for each of the N components. The two sidebands are present for each frequency present in the voiceband signal, so that, in general, spectral symmetry is not guaranteed for the composite signal.

5. Phase Intercept Distortion

Phase intercept distortion is θ_o in Equation (3), it is the frequency and time invariant component of phase shift in the received signal waveform. Under basic channel characteristics, measurement of the phase shift through a communications channel is difficult to perform. Even on a looped back connection through a communications network it is difficult to isolate phase intercept distortion from phase nonlinearity. Phase intercept distortion is not easily controlled on a carrier systems but it has no adverse effects on voice transmission. Designers of terminal modulating equipment for nonvoice signals have circumvented phase intercept distortion by frequency translation of baseband signals and derivation of the local demodulating carrier from received waveforms.

6. Frequency Shift

The f in the second term of Equation (3) is frequency shift, which may be generated in the following manner. Military communications systems frequently use single sideband carrier transmission facilities. In these systems the carrier frequency is not transmitted with the signal so that the signal may be demodulated with a locally generated carrier that differs slightly from the modulating frequency. This introduces a fixed frequency shift for all single frequencies by an amount equal to the difference between the modulating and demodulating frequencies. In some older carrier systems still in use, frequency shifts greater than 5 Hz have been observed due to relatively poor control of the difference between the modulating and demodulating carriers. In newer systems frequency shift is held to less than 1 Hz. Measurement of carrier frequency shift requires a tone source near 1000 Hz that is stable to at least one cycle per million. Both the transmitted and the received waveform's zero crossings must then be observed with frequency counters accurate to within ± 0.01 percent. The difference in zero crossing counts is frequency shift to the nearest 0.1 Hz.

Frequency counters are generally not balanced to ground. They may be sensitive to extraneous noise picked up on the test leads or from longitudinal currents on the line under test. To avoid disruption and insure repeatability of frequency measurements, it is desirable to place a 200-300 Hz bandpass filter centered at 1020 Hz just ahead of the counter.

7. Phase Jitter

The third term in Equation (3) represents all of the AC components of incidental PM, which cause the zero crossings of a voiceband signal to "jitter". Phase jitter measurements also include the disturbing effects of uncorrelated interference and quantizing noise. In fact phase jitter measurements should always be accompanied by a signal-to-noise measurement to ascertain what portion of the measurement is due to incidental PM and how much is noise. Figure 36 demonstrates jitter readings produced by quantizing noise on a time division multiplex system and Figure 37 illustrates the effect of uncorrelated white Gaussian noise on a typical phase jitter measuring set. A test set that would respond to pure phase modulation only and ignore noise is highly desirable but not is currently available.

The instantaneous phase of the received data signal is likely to jitter, typically, at rates of 180 Hz and below causing sidebands with magnitudes of approximately 18 dB below the level of the carrier. This is approximately 15 degrees peak-to-peak. This effect is primarily caused by ripple in the DC power supply appearing in the master oscillator of long haul carriers and being multiplied through many stages. Some phase jitter also occurs in short haul systems from incomplete filtering of image sidebands. Digital carrier systems also will exhibit jitter at certain input frequencies.

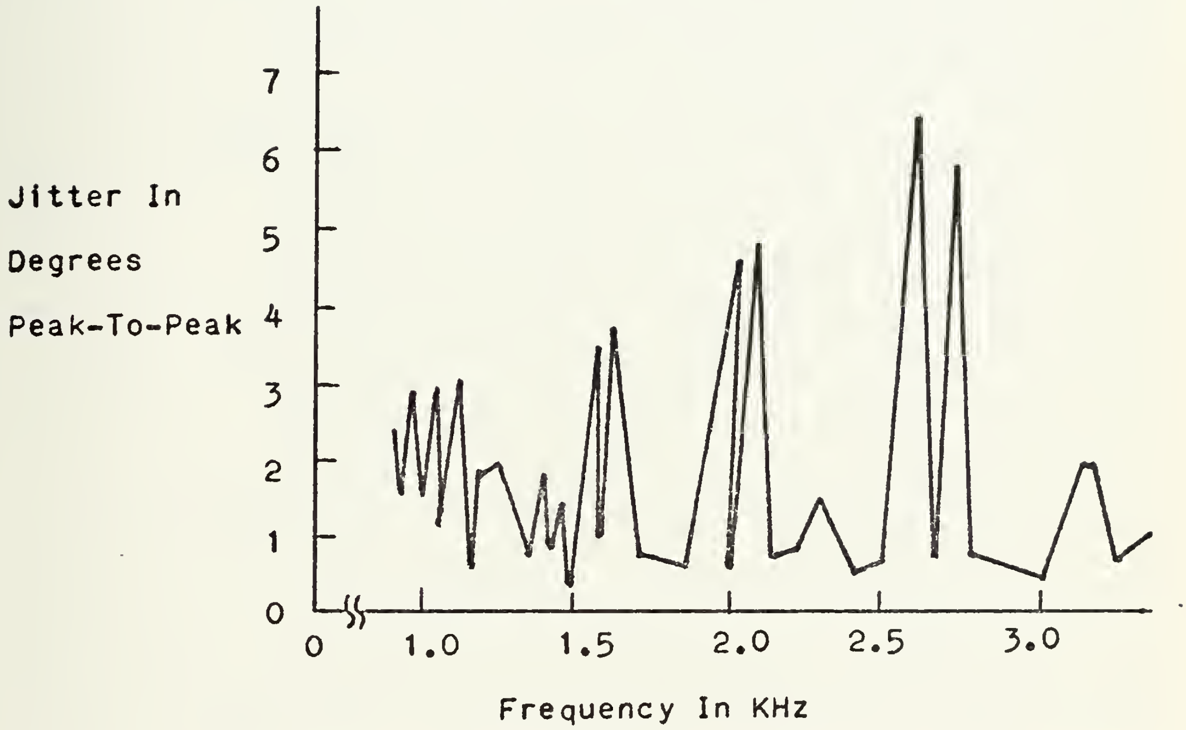


Figure 36- Typical Phase Jitter Measurement On A PCM System As A Function Of Carrier Frequency.

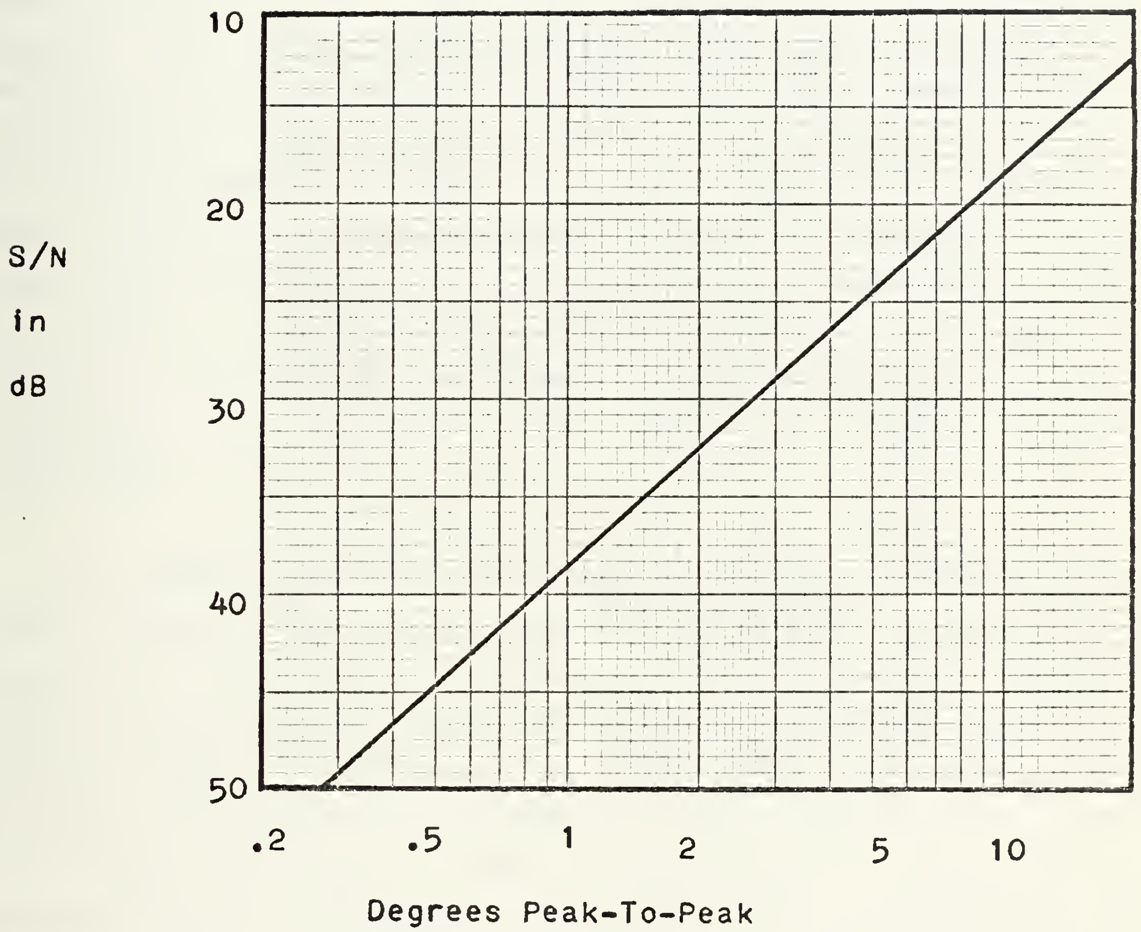


Figure 37- Typical Response Of Jitter Test Set To Noise
(3.3 KHz Of White Gaussian Noise)

Phase jitter measurements are made to assess the disturbing effect of undesired phase modulation upon received data signals. The peak-to-peak instantaneous phase deviations of the received carrier signal are measured, normally with respect to a local oscillator that is phase-locked to the long term average of the received signal. The phase-locked signal provides a jitter-free reference which is used as a basis for the phase jitter measurement. Since random noise can cause what would appear to be a significant amount of phase jitter, a C-notched noise measurement should always be made in conjunction with phase jitter measurements. Also, because quantizing noise can cause a significant phase jitter reading, care must be exercised in the choice of the carrier frequency and in the filtering to suppress the effect of noise on the measurement.

The most commonly found single-frequency components of phase jitter are 20 Hz (Ringing Current), 60 Hz (Commercial Power) and the second through fifth harmonics of these. Since the peak phase deviation caused by AC components of PM rarely exceeds 0.2 radians (low index phase modulation) only one pair of significant sidebands are produced for each sinusoidal component. Hence, a bandwidth of about 600 Hz centered about a carrier at or near 1020 Hz suffices to recover the major suspected sinusoidal PM without incurring large amounts of uncorrelated interference.

8. Incidental Amplitude Modulation

Incidental AM on communication channels takes the form of low index double sideband modulation of voice-band signals. Thus, referring to equation (3), m(t) is envelope modulation on the sinusoidal carrier of frequency f_o with received amplitude $A_o G(f_o)$. The incidental AM waveform may be expressed as follows:

$m(t) = \sum_{\alpha=1}^{M} m_a \operatorname{Cos}(2 f_a t+\phi_a)$. Where m_a, f_a, ϕ_a define the M-independent sinusoidal components of the amplitude modulation with low peak index ($\sum m_a < 0.2$). (The summation does not mean to imply to a Fourier series.) A single set of sidebands, $f_o + f_a$, are produced for each AM component but, except for their phase relationship to the carrier signal, these are indistinguishable from PM sidebands. Also, since incidental AM is low index, only small peak-to-peak excursions of carrier are evident and simple observation by oscilloscope or envelope detection makes the AM extremely difficult to distinguish from additive uncorrelated interference (especially hum and other single frequency interference). Hence, as was the case for incidental PM, band limiting and removal of other interfering modulation by a circuit analogous to the limiter for PM are implied.

Little concern has been expressed up to the present time over incidental AM on telecommunications channels. Its effect on voice transmission is negligible but with the advent of high speed (9600 bps) modems it may become a

parameter of interest. Some high speed modems show a sensitivity to incidental AM when it reaches a magnitude of about 10%. Almost no data exist to estimate the magnitude of this impairment.

9. Phase Hits And Gain Hits

Gain and phase changes that occur very rapidly may be encountered on telecommunciations channels. These are transient phenomena that might be thought of as components of the m(t) and o(t) of Equation (2). A phase or gain hit is defined as being a phase or gain change which lasts for a short period of time after which the signal switchs back to its original phase or gain. A phase or gain change is defined as one which occurs in the signal and it then remains at this new phase or gain for an indeterminate amount of time. Some of the more common causes of these phenomena are automatic switching to standby facilities or carrier supplies, patching out working facilities to perform routine maintenance, fades or path changes in microwave facilities and noise transients coupled into carrier frequency sources. The channel gain and phase (or frequency) shift may return to its original value in a short time or remain at the new values indefinitely.

Gain changes are typically detected by changes in the AGC circuit and phase changes by means of a phase locked loop. In order to provide protection against the detectors falsely operating on peaks of uncorrelated noise (impulse noise) a guard interval of 4 milliseconds should be designed into the peak indicating instrument. Unfortunately, such a guard interval will also effectively mask out true phase impulses shorter than 4 milliseconds that are not also accompanied by a peak amplitude excursion large enough to be detected by the threshold devices. This risk is considered justified at this time when the known relative frequencies

of phase jumps are compared with those for impulse noise.

Instruments used to measure gain and phase hits, as the rapid gain and phase changes are usually called, do so by monitoring the magnitude and phase of a sinusoidal tone. Hits are recorded and accumulated on counters with adjustable threshold levels. Gain hit counters typically accumulate events exceeding a threshold of 3 dB, although they do not distinguish an increase from a decrease of magnitude. Simılary, phase hit counters accumulate changes at thresholds from 10 to 45 degrees in 5-degree steps. They respond to any hits equal to or in excess of the selected threshold. A switch which removes the impulse noise blanking feature under the user's discretion may be desirable, when impulsive phase hit activity is suspected. As with the impulse noise counters previously discussed a controlled dead time of 140 ms should be built into the counters in order to obtain consistent readings with different types of sets.

A phase hit is defined as being a phase change which lasts for a short period of time after which the signal switches back to its original phase. A phase change is defined as one which occurs in the signal and it then remains at this new phase for an indeterminate amount of time. The following statistics on phase hits (and gain hits) were collected by monitoring five long-haul Bell System Communication channels for a total of about 70 hours. They are meant to be descriptive only and do not quantify the occurrence of these events in Bell System channels.

When the 70 hours was broken down into 15-minute intervals, it was observed that 70 percent of the intervals had no phase hits in excess of 22.5 degrees. The interval with the largest number of phase hits had 12 of them (see Figure 38). The distributions of the magnitudes of phase

hits and phase changes are shown in Figure 39. Both plots in Figure 39 start at 22.5 degrees. This was instrumentation limitation. Phase changes less than 22.5 degrees were not recorded. The distribution of the durations of phase hits is shown in Figure 40. The distribution shown is a conditional one. The phase hit duration had to be between 1.6 and 220 ms to be entered into the distribution. Most data modems will have accommodated the new phase before 220 ms and be operating error free again. This also was an instrumentation limitation. Note that approximately 13 percent of the phase hits are shorter than 4 ms in duration and would be missed by a phase hit counter which has the 4 ms guard interval in it.

Gain hits appear to be more common in the network than phase hits. The distributions of occurrences in 15-minute intervals are shown in Figure 41. Note, that at a 2-dB detection level, only 58 percent of the intervals had no gain hits. The largest number of gain hits observed in 15 minutes was 27. Also, as might be expected, the larger gain hits occur less frequently than the small ones. Figure 42 shows three distributions of the amplitudes of gain hits. Figure 43 shows distributions of the durations of gain hits for various detection levels. Note that 95 percent of the gain hits in excess of 10 dB last longer than 10 ms.

One serious problem with gain and phase hit counters occurs when a signal drop out on the order of 20 to 30 dB may occur and the background noise may simultaneously rise to a value near the original signal level. The phase locked loop will still recognize the desired signal but erroneous gain and phase hits are recorded due to the influence of the noise. This problem is partially mitigated by blanking the phase and gain hit counters during a drop out.

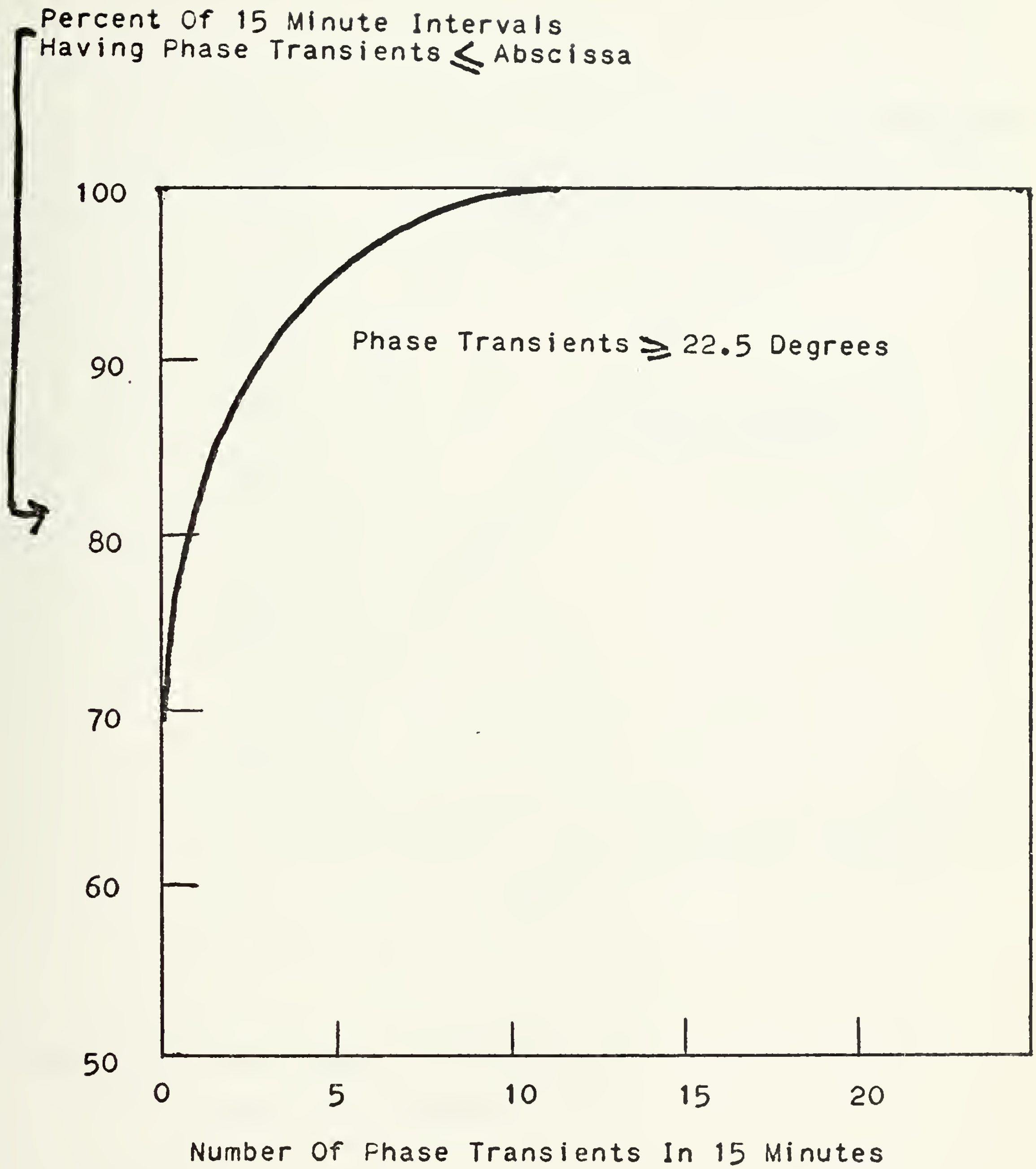


Figure 38- Distribution Of Phase Transient Counts In 15 Minutes

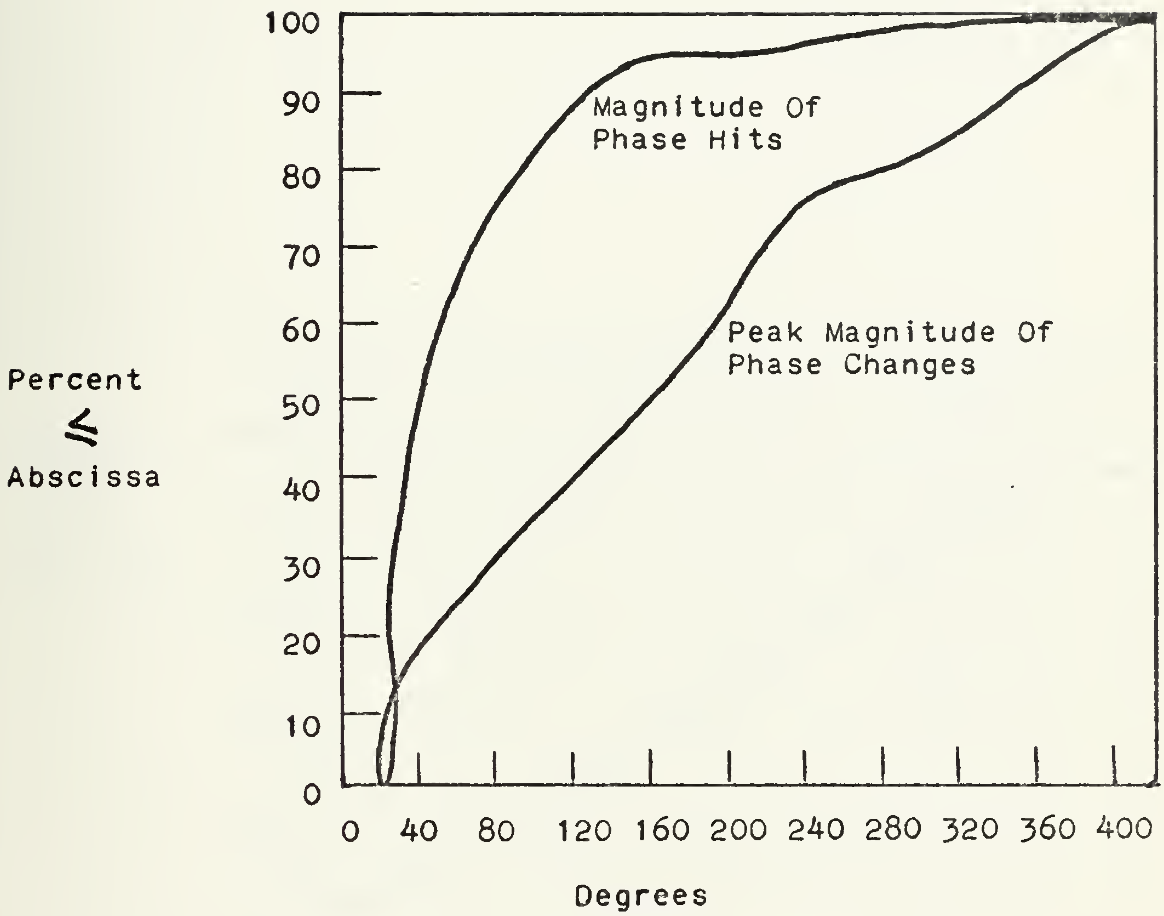


Figure 39- Distribution Of Magnitudes Of Phase Hits And Phase Changes.

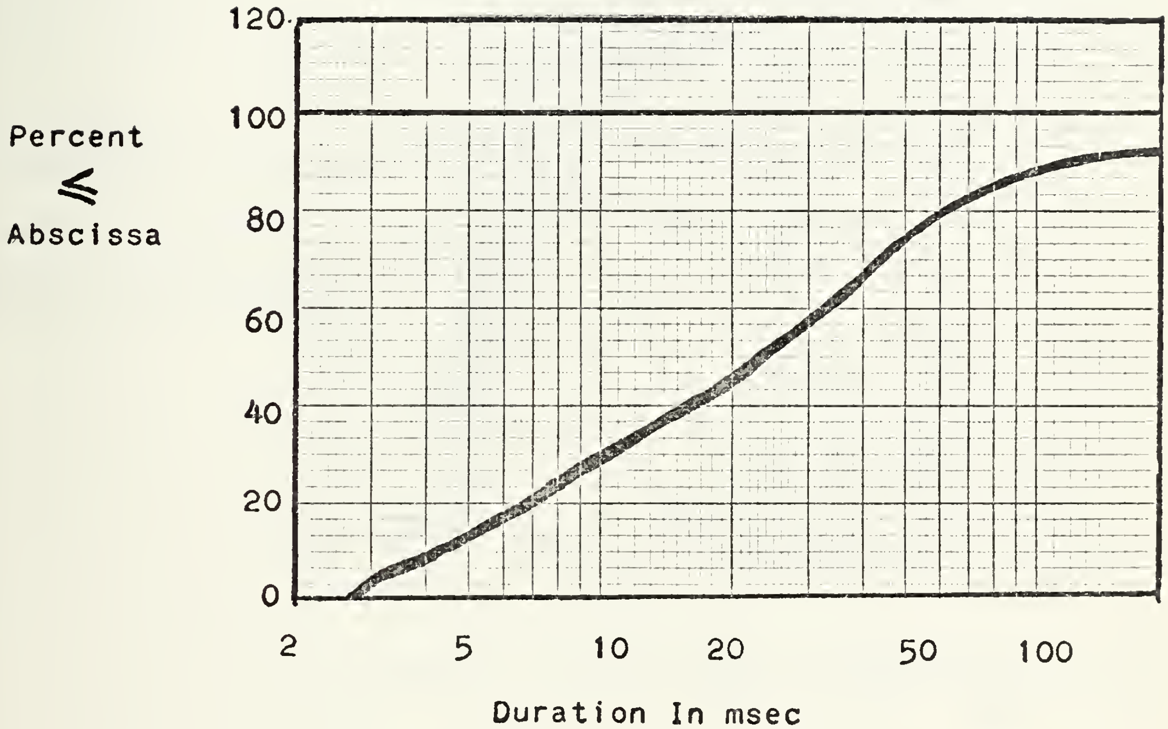


Figure 40- Distribution Of Duration Of Phase Hits Greater Than 22.5°, Given Durations Of 1.6 To 200 msec.

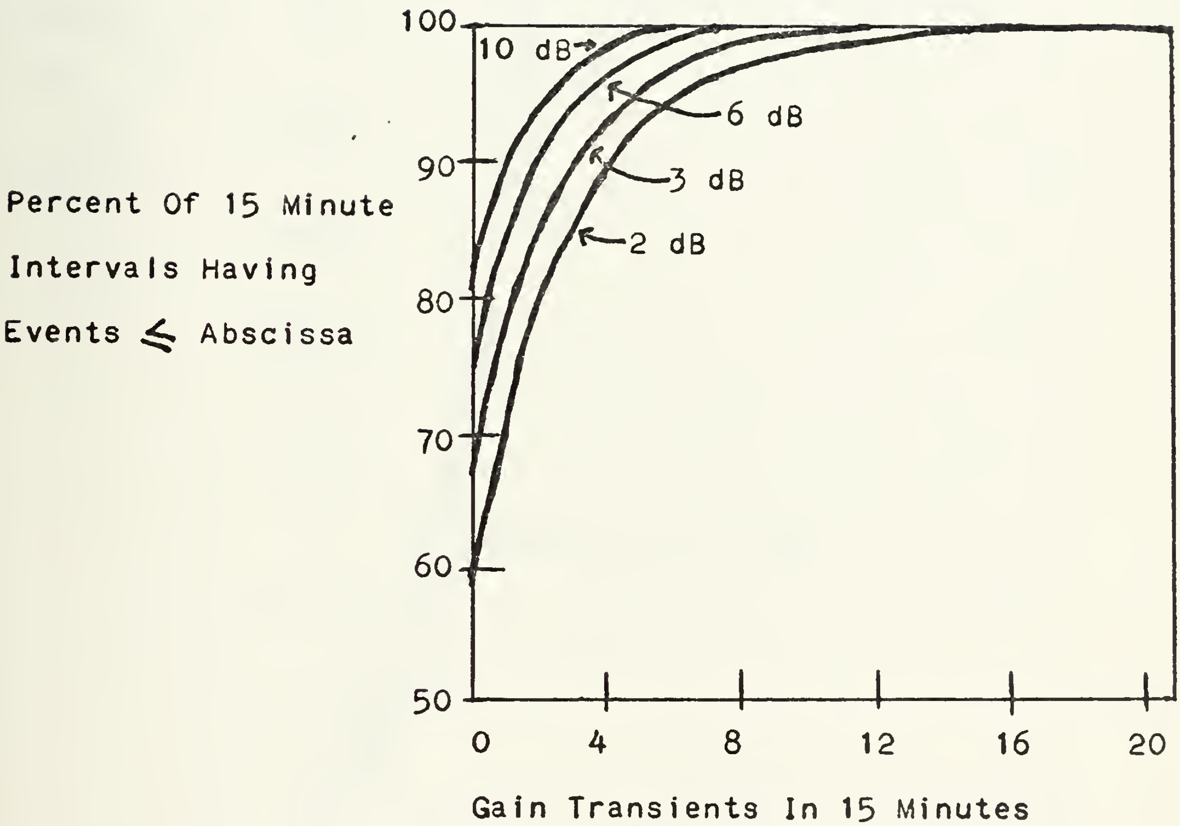


Figure 41- Distribution Of Number Of Gain Hits In 15 Minutes.

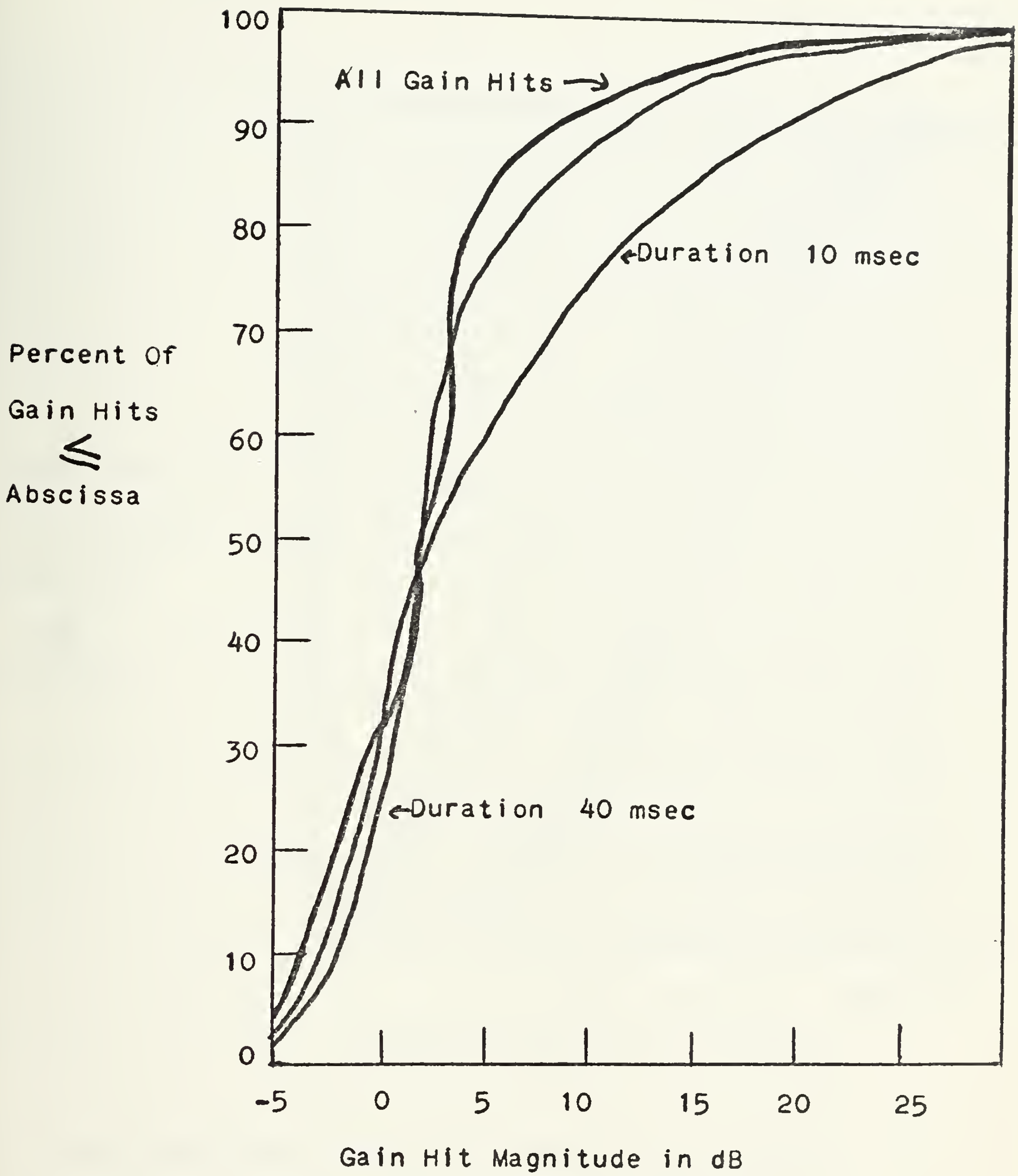


Figure 42- Distributions of Magnitudes Of Gain Hits.

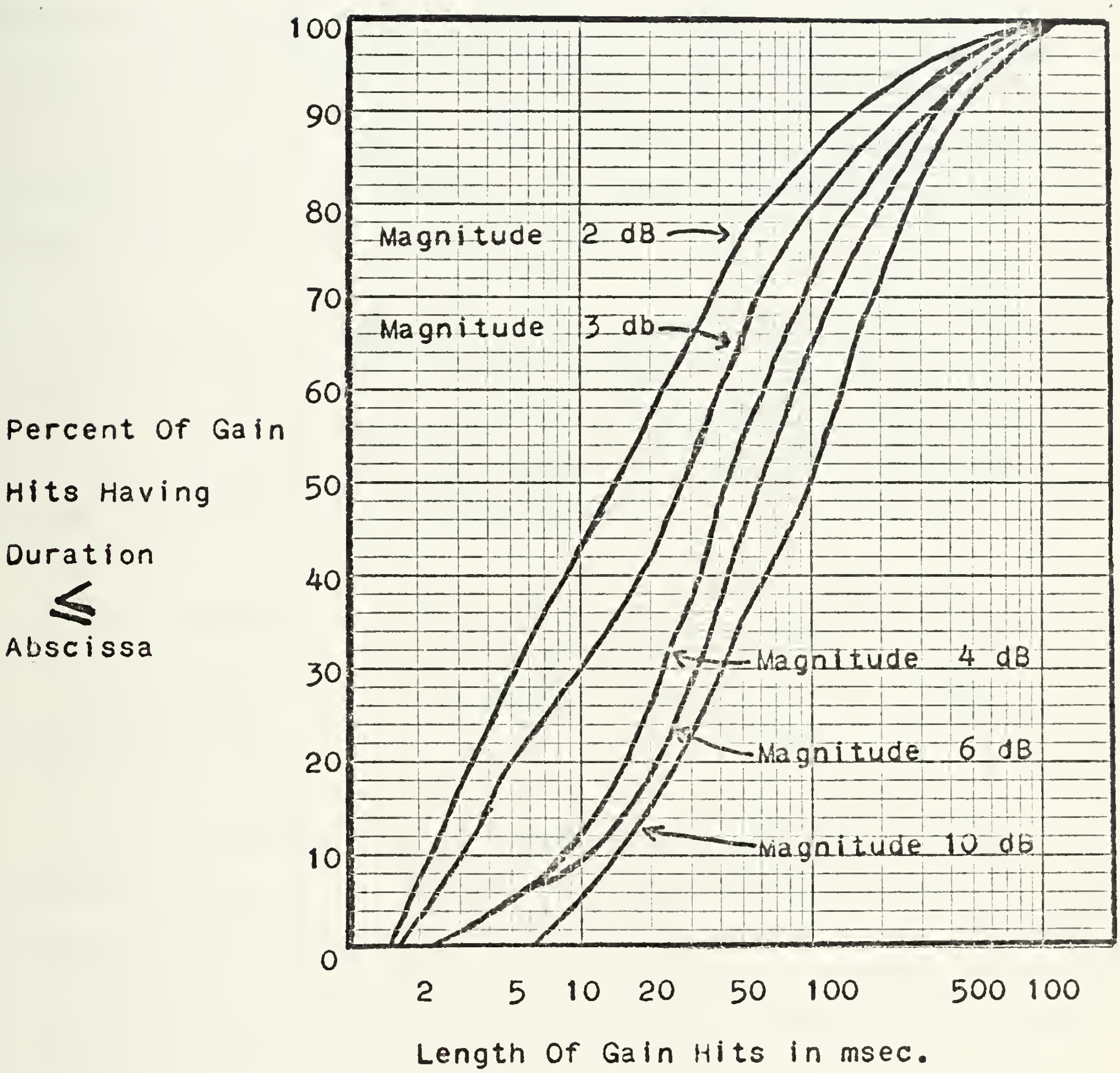


Figure 43- Distribution Of Length Of Gain Hits.

10. Dropouts

Given the A.G.C. range of most modems and from observations made on the channels mentioned previously it would appear that a most reasonable threshold for defining a dropout would be 12 dB. This would allow moderate decreases in signal level and still permit the modem to operate. Therefore, 12 dB is defined to be the threshold for calling large negative gain hits dropouts.

The data in Figure 43 can be extrapolated to estimate the minimum length of a 12 dB dropout. This occurs at about 10 ms. This then completes the definition of a dropout. It is any negative gain hit which is equal to or greater than 12 dB and lasts for at least 10 ms.

11. Summary

In summary the important parameters in regards to distortion are:

a. Envelope Delay Distortion: Envelope delay is defined as the derivative of the circuit phase shift (in radians) with respect to frequency (in radians per second). The deviation of this derivative at any frequency from its value at a perscribed frequency (usually 1800 Hz) is called envelope delay distortion.

b. Peak-To-Average Ratio (PAR): A measure of the transmission quality (mainly the phase response) of a channel for many data signals as derived from a peak-to-average ratio measurement of a particular test pulse.

c. Nonlinear Distortion: A measure of the second and

third order nonlinearities.

d. Frequency Shift: A fixed offset in each received frequency of a signal relative to the transmitted signal, due to differences in inserted carrier frequencies in receivers of transmission systems.

e. Phase Jitter: Undesired phase modulation on a received signal.

f. Phase Hits and Gain Hits: A phase or gain change which lasts for a short period of time after the signal switchs back to its original phase or gain.

g. Dropouts: Any negative gain hit which is equal to or greated than 12 dB and lasts for at least 10 ms.

h. Incidental Modulation: Any unwanted AM, PM, or FM imposed on the information carrying voiceband signal by a distrubing source other than itself. *

i. Phase Intercept Distortion: The frequency and time invariant component of phase shift in the received signal waveform. *

j. Incidental Amplitude Modulation: Low index double sideband amplitude modulation of voice signals. *

* At the present time, testing techniques and instrumentration have not been developed to measure these last three parameters easily or accurately.

V. PERFORMANCE MONITORING AND ASSESSMENT USING A TRANSMISSION IMPAIRMENT MEASURING SET

Most military technical control facilities have the equipment necessary to measure the significant parameters discussed in Section IV. The exception might be a bit error rate tester. Appendix B lists the equipment that would be required to perform the measurements. The analog test equipment would be able to measure the voice channels alone and isolate voice channel quality. The bit error rate tester would be needed to measure the digital variables but could not isolate problems in the data line due to malfunctions in the modem. As can be seen, with the many variables to be tested, and the amount of equipment required, the difficulty in setting up and measuring with all of this equipment would be enormous.

What is needed is a dedicated, task oriented, combination measuring set. Fortunately such a piece of equipment does exist commercially. The Hewlett-Packard 4940A Transmission Impairment Measuring Set (TIMS, see Figure 44) has been designed to do just one thing, that is to test the significant analog parameters of voiceband data communications networks in order to improve network performance. Specifically, the TIMS will test most of the parameters discussed in Section IV. Table III lists the parameters discussed in Section IV and indicates those tests that can be performed by the TIMS. The TIMS specifications are listed in Appendix C.

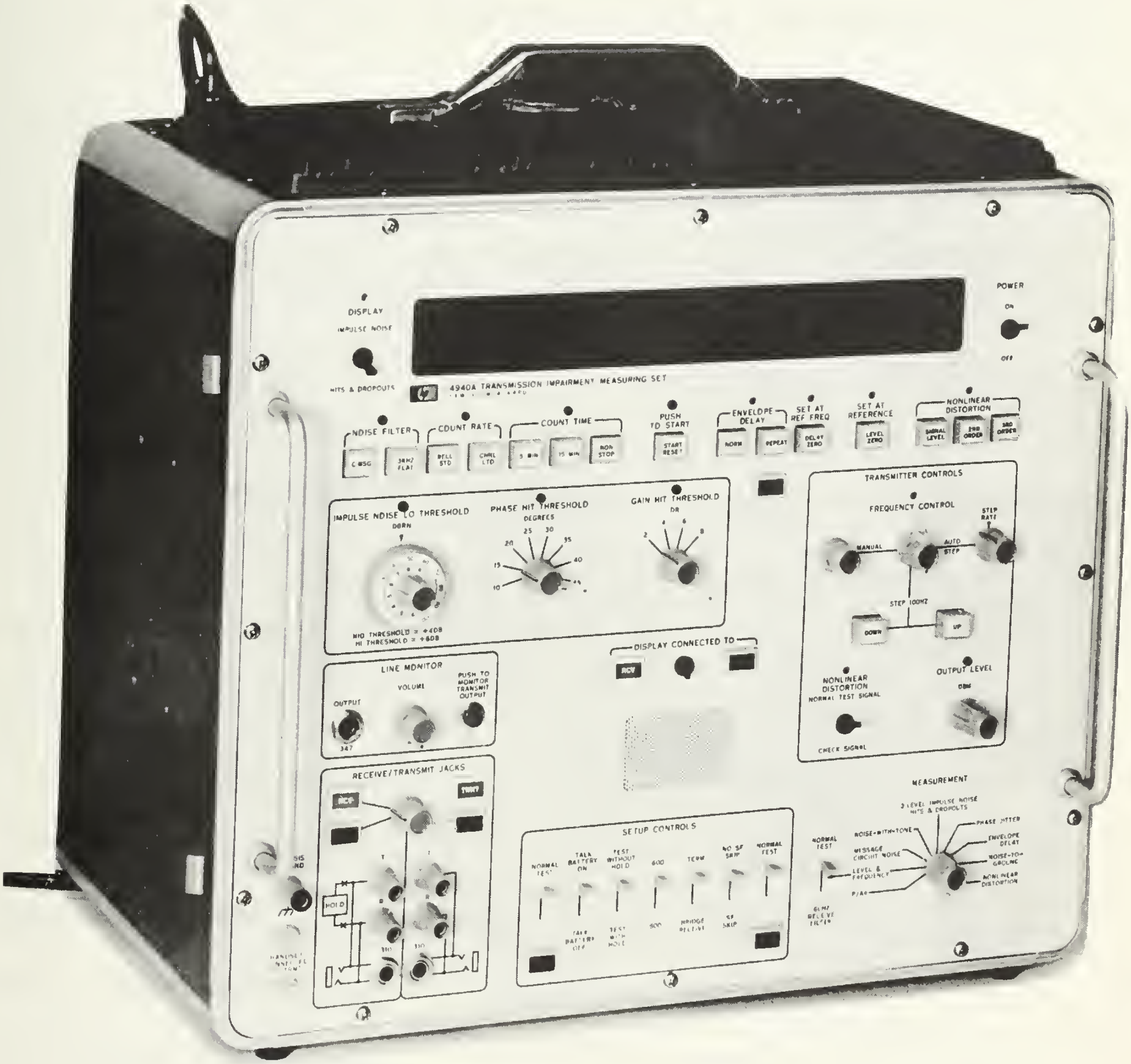


Figure 44- Hewlett Packard Model 4940A Transmission Impairment Measuring Set (TIMS)

TABLE III

Parameters Discussed in Section IV	Tests Performed by TIMS
Loss	yes
Return Loss	yes
Long Term Loss Variation	indirectly
Attenuation Distortion	yes
Baseband Loading	no
Message Circuit Noise	yes
C-Notched Noise (Noise-With-Tone)	yes
Impulse Noise	yes (3-level)
Single Frequency Interference	yes
Quantizing Noise	indirectly
Signal-To-Noise Ratio	indirectly
Envelope Delay Distortion	yes
Peak-To-Average Ratio	yes
Nonlinear Distortion	yes
Incidental Modulation	no
Phase Intercept Distortion	no
Frequency Shift	yes
Phase Jitter	yes
Incidental Amplitude Modulation	no
Phase Hits And Gain Hits	yes
Drop-outs	yes

Technical control personnel must be able to quickly measure parameters that affect transmission. Making all the measurements with seperate sets requires a lot of time and a technican with considerable experience. A voice channel can be thoroughly tested with the TIMS in a short amount of time. Trouble shooting data lines is much simpler, too. In one trip to the end of the data circuit, all parameters can be recorded to show all the data necessary to analyze and repair the problem.

The TIMS will permit technical controllers to isolate and test the analog section of the network. If this section is at fault, the common carrier can be contacted and given the information needed to begin repairs. If the analog line is not at fault, normal digital fault locating techniques can be used to identify a defective modem, communications terminal or other digital hardware. Using this procedure the terminal is contacted only when their equipment is at fault.

In the military communications system today, a high reliability data network is essential. A complete, straight forward routine testing of the analog channels is vital. The TIMS provides routine monitoring of parameters to be able to recognize slow degradations of channels and allow corrective action to be taken before the line actually fails.

The TIMS requires only one simple hook-up and needs only a short warm-up period. A reminder light system guides the technical controller through the measurements and monitors his switch selections for errors. Build in self checks allow the technical controller to quickly determine if his set is in working order.

All measurement values are displayed on high visibility digital readouts that simultaneously show three different parameters. Measurement annunciators alongside the readouts show the measurement unit displayed. Ranging, decimal placement, and polarity are all automatic.

A switch automatically programs the transmitter and receiver sections. If additional switch selections are necessary to complete a test, special indicators above the switches light up to guide the choice. If insufficient or erroneous selections are made, the set will blank out its display panel to prevent incorrect data from being recorded and will light a reminder lamp above the switch field causing the problem.

Level and frequency runs can be set up and logged in a short amount of time. The transmitter frequency can be stepped up or down from 204 Hz to 3904 Hz in 100 Hz increments. This can be done manually or automatically. Frequencies are offset 4 Hz to avoid interference problems with digital carriers. Level and frequency are displayed simultaneously at the transmit and receive ends on the digital display. Received level can be displayed either in absolute dBm or in dB relative to a desired reference. Attenuation distortion can be automatically calulated and displayed.

The same automatic frequency step controls can be used to make envelope delay runs. Measurements can be made using either the return reference or loop-around method. A no-loop indicator is provided to prevent erroneous readings when the envelope delay loop has been broken. Level, frequency, and delay are displayed simultaneously. The delay is shown clearly in microseconds with no calculation required. The TIMS will work end-to-end with all existing envelope delay sets which utilize the same technique.

Background message circuit noise can be tested in two ways: the traditional message circuit noise measurement with a quiet termination at the end of the circuit, or a noise-with-tone measurement can be used. The latter test measures the dynamic message circuit noise which is indicative of the actual background interference on a circuit with a data signal applied to it. The noise-with-tone measurement is made using a 1004 Hz tone to condition the voice channel circuits to their normal operating level. This tone is then filtered out and the residual message circuit noise is measured. A signal-to-noise ratio for the circuit can then be calculated with a simple subtraction. This is an important figure of merit for the channel. A noise-to-ground measurement is provided for to show how severe the common mode noise problem is on a particular circuit.

It is possible to simultaneously observe all the transients that cause data errors. By counting phase hits, gain hits, drop-outs and three levels of impulse noise at the same time, a more accurate analysis can be made of error causes and channel quality. All of these transients are totalled by the TIMS during the selected count time and stored in memory. The available count times are 5 minutes, 15 minutes and continuous. During the test and at the end of the count time, either the impulse noise totals or the hits and drop-out totals may be displayed from memory. The impulse noise threshold can be set to count impulses above 3 levels; low-level, mid-level (low +4 dB), and high-level (low +8 dB). The low impulse noise threshold can be selected at any level between 30 dBrn and 109 dBrn.

Phase hit and gain hit thresholds can also be set to count transients above a desired level. Drop-outs are automatically counted when the tone level drops more than 12 dB for longer than 10 milliseconds.

Peak-to-peak measurements of instantaneous phase deviations can be made by transmitting a 1004 Hz tone from one end of the line, and measuring the phase excursions of the received tone with a TIMS at the other end. The TIMS will display phase jitter, received frequency, and received level simultaneously. The TIMS is compatible with existing phase jitter test sets since it can receive test tones from 990 Hz to 1030 Hz.

Nonlinear distortion is a measurement especially relevant for transmission at higher data speeds utilizing multilevel loading. The TIMS utilizes an intermodulation distortion technique to measure nonlinear distortion. In addition, it used two pairs of tones to obtain consistent readings on networks with multiple sources of distortion and with digital carrier links. The test set will show second and third order distortion products readings either in dBm or directly in dB as a signal-to-distortion ratio. There is also incorporated into the set a check signal that allows correction for the influence of background noise on the distortion measurement.

The peak-to-average ratio (PAR) is useful to determine whether or not several parameters have changed since they were last recorded. Troubleshooting with the set in the PAR measurement mode will help to identify the direction of transmission with the worst characteristics. The PAR is a single number rating; a composite of the attenuation distortion, envelope delay distortion, nonlinear distortion, and message circuit noise on a channel. It is a figure indicative of the degradation that a data signal would undergo over a channel. The PAR test waveform is a synthesis of 16 seperate weighted frequencies from 140 Hz to 3900 Hz designed to approximate the spectrum of a data set.

Some other features of interest of the TIMS are :

Internal 600 ohm and 900 ohm terminations for both receiving and transmitting modes. It can also be connected in a bridging configuration in the receiving mode.

A talk circuit on the line under test can be established directly through the set. Connections are provided for connecting a lineman's handset for dialing and talking. Talk battery is provided so that communications can be set up on dry lines. A holding circuit can also be switched in to allow tests to be run on a dialed-up line.

A built-in speaker permits the monitoring on either the transmit or receive circuit. This allows the technical controller to make a listening analysis of circuit phenomena which cannot be analyzed quantitatively, such as the listening test for single frequency interference.

A sample TIMS transmission line parameter record is shown in Figure 45. All the data indicated on the record can be measured and logged in less than 30 minutes. Utilizing separate instruments to measure all of these parameters would take two experienced technicans over two hours.

As can be seen the transmission impairment measuring set can perform the functions of approximately 20 seperate communications test sets now in use by military technical control personnel. It can perform the measurements required for system analysis in a much shorter period of time than is now required. These measurements, intelligently analyzed, will allow technical controllers to monitor and assess system performance, so as to be able to predict failures.

The advantages of the TIMS over the equipment now in use are:

1. Less expensive (approximately 2 to 1 savings).

2. Shorter set-up and measurement time (approximately a 4 to 1 reduction).

3. Less technical training required.

4. Less space required (approximately 10 to 1 reduction in amount of space occupied, this is particularly important for mobile technical control facilities).

5. Up-to-date measurement techniques available for measurement of modern system parameters.

CIRCUIT DESCRIPTION 4-WIRE 3002

TIMS TRANSMISSION LINE PARAMETER RECORD

TEST NO. 3

TEST DATE 12/12/73

REASON FOR TEST ROUTINE VERIFICATION OF CHANNEL CHARACTERISTICS

FROM PALO ALTO TO ATLANTA

1. LEVEL & CONNECTION VERIFICATION
   - 1000 HZ RCV -16.2 DBM
   - TRMT TLP (FROM SIDE) +7
   - RCV TLP (TO SIDE) -3
   - 60 HZ FILTER NO
   - FREQ SHIFT NO
2. LOSS 10.2 DB
3. P/AR 83 UNITS
4. NOISE

	C MSG	3 KHZ
MSSG CRCT	36 DBRN	33 DBRN
WITH TONE	37 DBRN	44 DBRN
S/N RATIO	37 DB	30 DB
TO GROUND	67 DBRN	70 DBRN

DESCRIBE NOISE: NO POPPING NO SINGLE FREQUENCY INTERFERENCE

5. PHASE JITTER 6.5 °
6. NONLINEAR DSTRTN

	2 ND	3 RD
TEST	41 DB	38 DB
CHECK	44 DB	46 DB
CORRECTED	44 DB	39 DB

7. LEVEL/FREQUENCY

300	+3.4 DB	2000	+.6 DB
500	2.1 DB	2200	1.0 DB
600	1.3 DB	2400	2.0 DB
800	.8 DB	2500	2.7 DB
1000	0 DB	2600	3.1 DB
1200	-.6 DB	2700	3.8 DB
1400	-.4 DB	2800	4.8 DB
1600	+.1 DB	3000	6.0 DB
1800	+.4 DB	3200	9.9 DB

8. ENVELOPE DELAY

500	915 µS	2000	-10 µS
600	754 µS	2200	103 µS
800	331 µS	2400	220 µS
1000	175 µS	2500	412 µS
1200	118 µS	2600	580 µS
1400	37 µS	2700	730 µS
1600	25 µS	2800	945 µS
1800	0 µS	3000	1122 µS

9. IMPULSE NOISE, PHASE HITS, GAIN HITS, DROP OUTS (BELL STD, C MSG)
   - 15 CNT TIME

68 DBRN	LO COUNTS	1
+4DB	MID COUNTS	5
+8DB	HI COUNTS	0
15 °	PHASE HITS	4
	DROP OUTS	0
2 DB	GAIN HITS	1

FROM ATLANTA TO PALO ALTO

1. LEVEL & CONNECTION VERIFICATION
   - 1000 HZ RCV -16.1 DBM
   - TRMT TLP (FROM SIDE) +7
   - RCV TLP (TO SIDE) -3
   - 60 HZ FILTER NO
   - FREQ SHIFT NO
2. LOSS 10.1 DB
3. P/AR 72 UNITS
4. NOISE

	C MSG	3 KHZ
MSSG CRCT	37 DBRN	41 DBRN
WITH TONE	39 DBRN	47 DBRN
S/N RATIO	35 DB	27 DB
TO GROUND	55 DBRN	73 DBRN

DESCRIBE NOISE: NO POPPING NO SINGLE FREQUENCY INTERFERENCE

5. PHASE JITTER 5.8 °
6. NONLINEAR DSTRTN

	2 ND	3 RD
TEST	50 DB	47 DB
CHECK	57 DB	53 DB
CORRECTED	51 DB	48 DB

7. LEVEL/FREQUENCY

300	+6.3 DB	2000	-.9 DB
500	+3.3 DB	2200	-.1 DB
600	+1.9 DB	2400	+.9 DB
800	.4 DB	2500	+2.4 DB
1000	0. DB	2600	4.3 DB
1200	-.3 DB	2700	6.5 DB
1400	-.5 DB	2800	8.5 DB
1600	-.8 DB	3000	11.7 DB
1800	-1.2 DB	3200	DB

8. ENVELOPE DELAY

500	1170 µS	2000	-18 µS
600	910 µS	2200	+12 µS
800	650 µS	2400	119 µS
1000	290 µS	2500	360 µS
1200	110 µS	2600	758 µS
1400	78 µS	2700	1120 µS
1600	29 µS	2800	1390 µS
1800	0 µS	3000	µS

9. IMPULSE NOISE, PHASE HITS, GAIN HITS, DROP OUTS (BELL STD., C MSG)
   - 15 CNT TIME

68 DBRN	LO COUNTS	9
+4DB	MID COUNTS	4
+8DB	HI COUNTS	1
15 °	PHASE HITS	2
	DROP OUTS	0
2 DB	GAIN HITS	1

Figure 45- TIMS Transmission Line Parameter Record

VI. CONCLUSIONS

The purpose of this thesis was to determine and analyze the specific parameters that are required for performance monitoring and assessment of military communications systems by military technical controllers. These parameters if properly measured and analyzed will allow technical controllers to anticipate difficulities and permit appropriate corrective action before actual failure of the communications system. The following parameters were arrived at:

1. Loss: End-to-end circuit attenuation, usually measured at 1000 Hz.

2. Return Loss: A measure of the mismatch between the actual circuit impedance as compared to a nominal impedance defined for ideal circuits.

3. Long Term Loss Variations: Changes in the loss of a circuit due to aging of components, changes in physical makeup, and temperature variations.

4. Attenuation Distortion: Loss deviation (1000 Hz reference) over the range of frequencies of interest. This includes static and dynamic frequency response and bandwidth.

5. Baseband Loading: The result of high levels in the baseband of multiplexed circuits. These high levels in the baseband raise the loading level to the modulator and can cause severe intermodulation noise on the channels.

6. Message Circuit Noise: Background noise measured through a weighting network.

7. C-Notched Noise: A measure of the background or impulse noise measured through a weighting network when a holding tone, usually 2805 Hz at -10 dBm0, is being transmitted over the system under test. The tone is blocked at the measuring set by a notch filter.

8. Impulse Noise: Indicated by the number of noise bursts exceeding a selected voltage threshold.

9. Single Frequency Interference: Spurious tones present on the channel in addition to the desired signal.

10. Quantizing Noise: Signal correlated noise generally associated with the quantizing error introduced by analog-digital and digital-analog conversions in digital transmission systems.

11. Envelope Delay Distortion: Envelope delay is defined as the derivative of the circuit phase shift (in radians) with respect to frequency (in radians per second). The deviation of this derivative at any frequency from its value at a perscribed frequency (usually 1800 Hz) is called envelope delay distortion.

12. Peak-To-Average Ratio (PAR): A measure of the transmission quality (mainly the phase response) of a channel for many data signals as derived from a peak-to-average ratio measurement of a particular test pulse.

13. Nonlinear Distortion: A measure of the second and third order nonlinearities.

14. Frequency Shift: A fixed offset in each received frequency of a signal relative to the transmitted signal, due to differences in inserted carrier frequencies in

receivers of transmission systems.

15. Phase Jitter: Undesired phase modulation of a received signal.

16. Phase Hits and Gain Hits: A phase or gain change which lasts for a short period of time after the signal switchs back to its original phase or gain.

17. Dropouts: Any negative gain hit which is equal to or greated than 12 dB and lasts for at least 10 ms.

In conclusion it must be emphasized that because the military communication systems of today are increasingly more complex and sophisticated, it is necessary that the individuals who control the operation of these systems have a thorough knowledge of the parameters that are required to adequately monitor and assess system performance. The existing go or no-go philosophy of technical control indicates that the knowledge of these basic system parameters apparently does not exist. It is not enough to prove that a circuit has failed. The technical controller must be able to predict any system degradation or impending failure. Knowledge of the basic parameters discussed in this thesis coupled with the ability to measure and analyze these parameters will allow the technical controllers to monitor and assess system performance so as to be able to predict failures. It can also be concluded that a dedicated, task oriented, combination measuring set such as the HP 4940A Transmission Impairment Measuring Set described in this thesis will perform the measurements required for performance monitoring and assessment in a highly efficient manner.

APPENDIX A

BASIC CONCEPT OF MILITARY TECHNICAL CONTROL

A. DEFINITION AND FUNCTIONS OF MILITARY TECHNICAL CONTROL

Technical control is a term designating the functions of technical direction, coordination, technical supervision of transmission media and equipment, quality control, communications service restoral, and status reporting required in order to provide effective communications services to the users. A Technical Control Facility (TCF, see Figure 46) is a station that functions as the point of interface between the transmission elements of the system, interfaces users with the system and has the physical, electrical, and manpower capabilities to perform the following functions:

1. Exercise technical direction, coordination, and supervision over transmission links, supergroups, groups, channels, circuits, interfacing equipment appearing in the TCF, remote transmitter sites, receiver sites, radio relay sites and switching-relay sites, as well as those extension communications facilities provided by or to all directly connected users.

2. Restore disrupted service to users via any remaining available means on a predetermined restoration priority and near real-time basis.

3. Perform quality control checks and tests on all channels, circuits, and equipment appearing in the TCF. Exercise technical supervision over the performance of quality control checks and tests on all transmission links, supergroups, and groups entering or leaving the TCF.

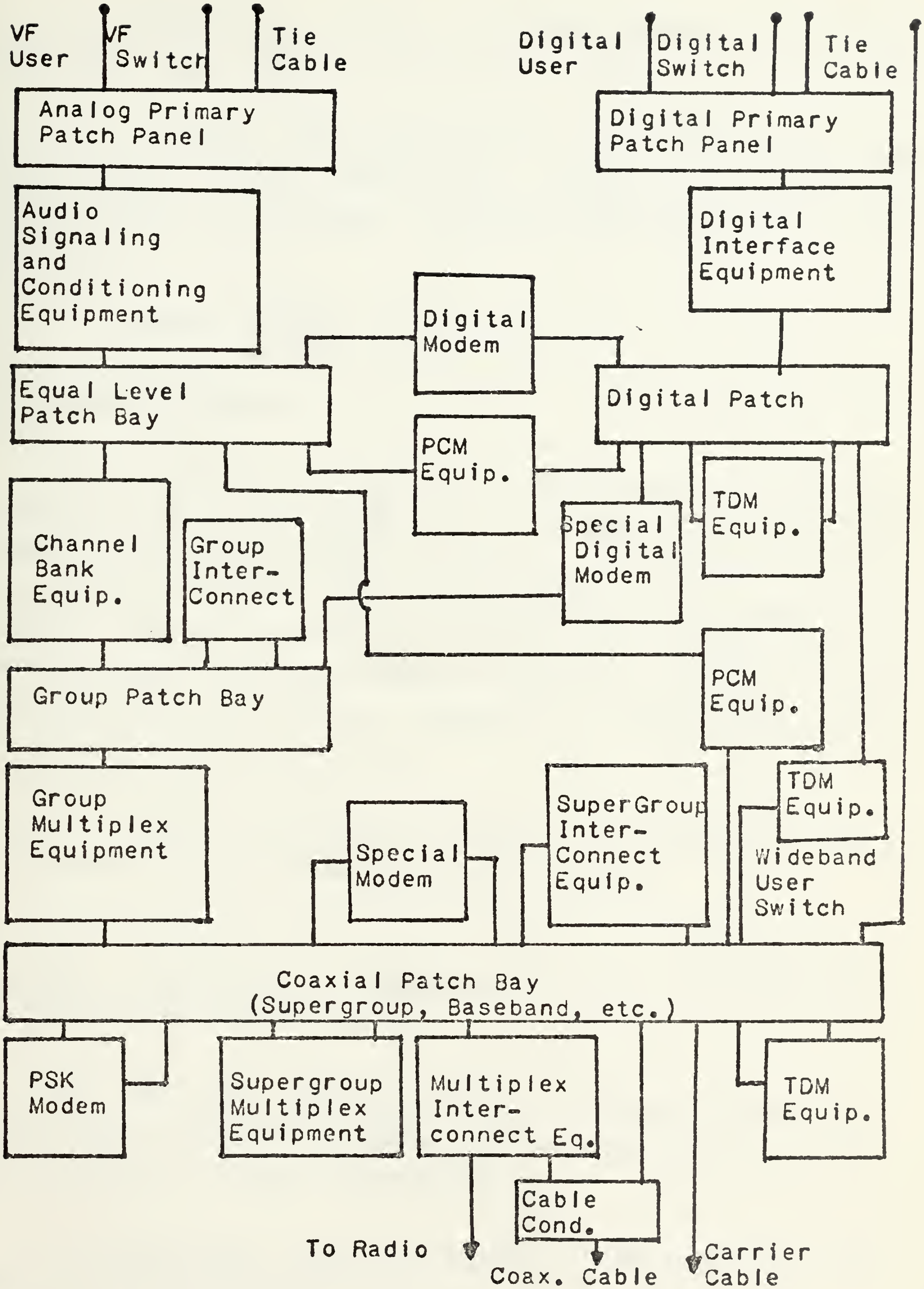


Figure 46- Technical Control Facility Block Diagram

4. Activates, changes, and deactivates circuits in accordance with Circuit Engineering Orders.

5. Report to appropriate agency the status of transmission links, supergroups, groups, channels, and circuits for operational direction, management and record purposes.

6. Isolate communication faults.

B. QUALITY CONTROL

The technical controller maintains quality and continuity of communications by taking appropriate action in accordance with established technical control procedures and the station standard operating procedures to prevent or correct conditions adversely affecting the operation of the communications circuits under his control. He is to provide the following preventive measures to assure quality and continuity of communications.

1. Perform operational tests on circuits to ensure high quality operation and maximum efficiency.

2. Periodically verify quality of communications circuits, channels, and equipment by use of appropriate communications test equipment.

3. Use test devices, such as signal generators, meters, and distortion analyzers in conjunction with monitor aids to analyze circuit conditions.

4. Determine over-all circuit conditions from test results and observations.

5. Determine appropriate action to be taken based on analysis of test results.

6. Apply established procedures using voice or data (teletypewriter) order wires, in coordination with local and distant facilities, to effect necessary operational adjustments on carriers and associated equipment.

7. Direct HF radio frequency changes.

8. Place additional authorized circuits in operation as required by traffic load.

9. Coordinate utilization of cryptographic or other security equipment.

10. Coordinate with local using agencies, other military services, and commercial communications organizations in matters relating to circuit performance, capabilities and utilization.

11. Maintain required logs and records and provide data for preparation of pertinent reports.

12. Maintain continuous knowledge of facilities available for rerouting.

C. CORRECTIVE ACTION

The technical controller must also ascertain and initiate appropriate action to correct conditions affecting efficiency and continuity of communications as follows:

1. Coordinate with local facilities and distant technical controllers and conduct tests to locate source of trouble.

2. Participate in such tests by operating measuring and test devices.

3. Evaluate test results and factors involving interruptions, failures, or disturbances resulting from such situations as atomspheric storms, unpredictable ionospheric changes, interference from other stations, and equipment failures.

4. Ascertain in conjunction with local facilities and distant technical controllers appropriate corrective action, such as substitution of equipment, frequency changes, or other measures, such as establishing reroutes.

5. Record all time lost or inoperative circuits and other pertinent information for use in analysis of circuit failures.

6. Report all outages in accordance with perscribed procedures.

7. Coordinate with maintenance personnel pertaining to equipment malfunctions.

D. SKILLS AND KNOWLEDGE REQUIRED

The skills and knowledge deemed necessary for a technical controller are:

1. Must know and be proficient in application of technical control procedures.

2. Must know available reroutes and must be familiar with equipment available for use in restoring circuits.

3. Must know current communication principles and

concepts.

4. Must know operating characteristics, capabilities, limitations , and use of controlled communications equipment to include all transmission media used under his control.

5. Must know functional principles and application of various types of test devices, such as meters, oscilloscope, signal generators, and distortion analyzers.

6. Must know function and application of hand tools used in communicators electronics work.

7. Must know electrical and electronic fundamentals to include theory of operation of communications equipment.

8. Must be able to read and interpret equipment and patch panel schematic diagrams, and trunk and circuit layout records.

9. Must know principles of radio propagation and be able to interpret HF radio frequency propagation charts and forecast data.

10. Must be able to isolate cause of interference and recognize various types of circuit troubles.

11. Must know basic priciples of circuit conditioning and the applications of conditioning equipment.

12. Must know characteristics of fixed plant cable and wire line transmission theory.

13. Must know principles and application of wideband data and carrier equipment, and systems.

14. Must be able to interpret measurements indicated by test devices.

APPENDIX B

TEST EQUIPMENT REQUIRED TO MEASURE PARAMETERS

Table IV contains a listing of the test equipment necessary to manually measure the parameters described in Section IV. This list contains those equipments that should presently be available to every manual technical control facility. The equipment listed is the perferred model. There are many alternate models of older equipment which may be substituted to perform the measurements.

The list was derived from DCAC 310-70-1 and NAVCOMMCOM Instruction 2300.13. Figure 47 is an illustration of one possible cabinet configuration for a Performance Monitoring And Assessment Test Bay.

TABLE IV

TEST EQUIPMENT REQUIRED TO MEASURE PARAMETERS

1. AC Voltmeter/Audio Oscillator CAQI-3550B

2. Frequency Counter AN/USM-207

3. Oscilloscope AN/USM-296

4. Level/Noise Measuring Set (tones/voice) CDDD-TTS-85CR

5. Audio Level Meter (voice) TS-2778/USM

6. Envelope Delay Measuring Set TS-2395/G

7. Selective Noise Measuring Set CAQI-302A

8. Impulse Noise Measuring Set CDDD-TTS-58B

9. Transmission Measuring Set Type 12-B (Daven)

10. X-Y Recorder CAQI-7000A/AM

11. Log Convertor (for X-Y Recorder) CAQI-7560A

12. Signaling Test Set AN/TSM-86 (Lenkurt 26600)

13. Electronic Counter HP-5221A

14. Spectrum Analyzer CAQI-140B

15. Jitter Meter Hekiman 45

16. Digital Data Signal Generator SG-885(P)/USM

17. Data Analysis Center AN/CGM-15

18. Data Error Rate Test Set

19. Signal Generator AM/URM-25

20. PAR Meter

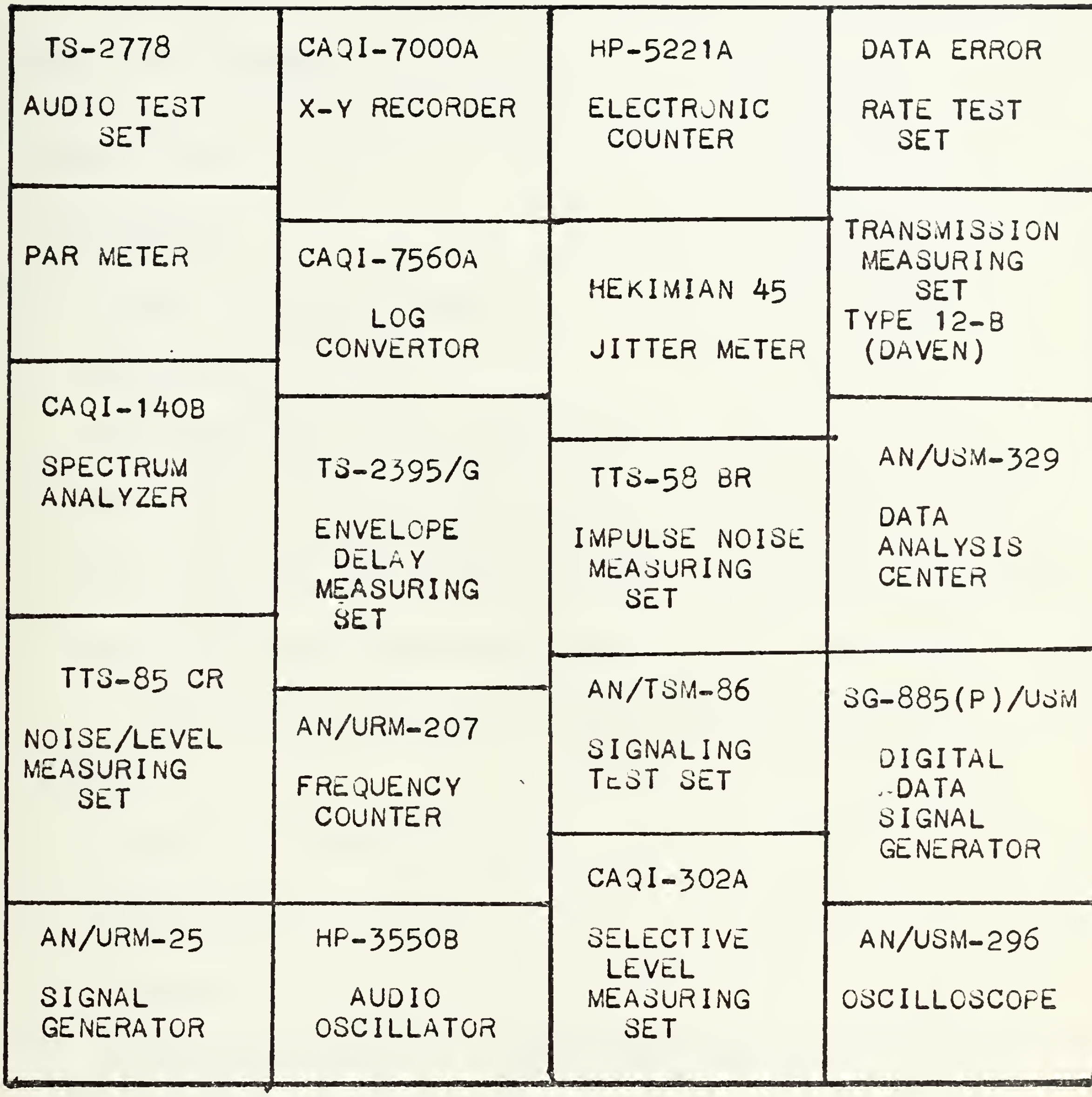


Figure 47- Performance Monitoring and Assessment Test Bay

APPENDIX C

TRANSMISSION IMPAIRMENT MEASURING SET SPECIFICATIONS

LEVEL AND FREQUENCY

Transmitter

Frequency

Range : 200 Hz to 3904 Hz

Resolution: 1 Hz

Accuracy: ±0.5 Hz in Step Modes

Manual Mode: Continuously adjustable; 200 Hz to 3900 Hz

Step 100 Hz Mode: 204 Hz to 3904 Hz in 100 Hz increments

Auto Step Mode: Repeatedly steps up and then back (in 100 Hz increments) from 204 Hz to 3904 Hz

Level

Range: +10 dBm to-40 dBm

Resolution: 0.1 dBm

Accuracy: ±0.1 dB

Frequency Response: ±0.1 dB from 200 Hz to 3904 Hz

Harmonic and Spurious Signals:

Total Harmonic Distortion to 10 kHz: Greater than 50 dB below fundamental

Spurious Signals to 10 kHz: Greater than 50 dB below fundamental

Background Noise to 10kHz: Less than -90 dBm

Receiver

Frequency

Range: 200 Hz to 4100 Hz

Resolution: 1 Hz

Accuracy: ±0.5 Hz

Level

Range: +10 dBm to-40 dBm

Resolution: 0.1 dBm

Accuracy: ±0.1 dB at 1 kHz from 0 to-15 dBm; ±0.2 dB over full frequency and level range

Detector: Full wave average

Bandwidth (3 dB points): Nominal 20 Hz to 10 kHz

60 Hz Input Filter (switch selectable): Nominal 25 dB loss at 60 Hz; 4 dB loss at 180 Hz

PHASE JITTER

Transmitter

Frequency: 1004 Hz fixed

Level

Range: + 10 dBm to-40 dBm

Resolution: 1 dBm

Background Noise Harmonics and Spurious Signals:

Refer to Level and Frequency

Receiver

Frequency Range: 990 Hz to 1030 Hz

Level

Range: +10 dBm to -40 dBm

Resolution: 1 dBm

Jitter

Range: 0.2 to 25 degrees

Resolution: 0.1 degree

Accuracy: ±5% of reading ±0.2 degrees

Detector: Peak to peak

Bandwidth (3 dB points): Nominal 20 Hz to 300 Hz

MESSAGE CIRCUIT NOISE

Transmitter

Quite Termination

Receiver

Weighting Filters: C-Message and 3 kHz flat

Detector: Quasi RMS

Range: +10 dBrn to +100 dBrn

Resolution: 1 dBrn

Accuracy: ±1 dB

3-LEVEL IMPULSE NOISE, HITS AND DROPOUTS

Transmitter

Frequency: 1004 Hz fixed

Level

Range: +10 dBm to-40 dBm

Resolution: 1 dBm

Background Noise, Harmonics and Spurious Signals:

Refer to Level And Frequency

Receiver

Received Tone Level: 0 dBm to-40 dBm

Notch Filter: Greater than 50 dB rejection

from 995 Hz to 1025 Hz

Count Interval: 5min., 15 min. or continuous

Impulse Noise Threshold Range:

Lo: 30 dBrn to 109 dBrn

Mid: 4 dB above Lo (Max 109 dBrn)

Hi: 8 dB above Lo (Max 109 dBrn)

Impulse Noise Threshold Accuracy: ±1 dB

Impulse Noise Count Range:

Lo: 0 to 19,999

Mid: 0 to 9,999

Hi: 0 tc 1,999

Count Bate 7 counts per second max. ±5%

or channel limited

Gain Hit Thresholds: 2, 4, 6, and 8 dB

Phase Hit Thresholds: 10 to 45 degrees in

5 degree increments

Hit Guard Interval: Nominal 4ms

Dropout-Level Threshold: Fixed; greater

than 12 dB decrease in level

Dropout Duration: Nominal 10 ms or greater

Phase Hit Counts: 0 to 19,999

Dropout Ccunts: 0 to 9,999

Gain Hit Ccunts: 0 to 1,999

Display: All six phenomena are

counted simultaneously and stored.

A switch selection displays either the Impulse Noise counts or the Hits and Dropout ccunts during or after the measurement period.

NOISE-TC-GRCUND

Transmitter

Quiet Termination

Receiver

Range: 40 dBrn to 130 dBrn

Input Circuit: 600 or 900 across the line:

100 k to ground

For other specifications refer to

Message Circuit Noise

ENVELOPE DELAY

Transmitter

Frequency

Range: 300 Hz to 3904 Hz

For other Frequency Specifications see

Level And Frequency

Modulation Frequency: 83 Hz

Level

Range: +10 dBm to -40 dBm

For other level specifications see Level And Frequency

Harmonic and Spurious Signals to 10 Hz:

Greater than 46 dB below main signal power.

Background Noise to 10 kHz: Less than -90 dBm

Receiver

Frequency Range: 300 Hz to 3904 Hz

For other Frequency specifications see

Level And Frequency

Level Range: +10 dBm to -40 dBm

Carrier Level Accuracy: ±.25 dB

For other level specifications see

Level And Frequency

Delay

Range: -3,000 s to +9,000 s

Resolution: 1 s

Accuracy (back to back):

±10 s from 600 Hz to 3904 Hz

±30 s from 300 Hz to 600 Hz

Minimum Signal to Noise Ratio (or stated accuracy):

20 dB, 3 kHz flat weighting

NOISE-WITH-TONE

Transmitter

Frequency: 1004 Hz fixed

Level

Range: +10 dBm to -40 dBm

Resolution: 1 dBm

Background Noise, Harmonics and Spurious Signals:

Refer to Level And Frequency

Receiver

Notch Filter: Greater than 50 dB rejection from 995 Hz

For other specifications refer to Message Circuit Noise

NONLINEAR DISTORTION

Transmitter

Test Signal Type: Two pairs of tones centered at 860 Hz and 1380 Hz

Level

Range: 0 dBm to-40 dBm

Resolution: 1 dBm

Receiver

Level

Range: +10 dBm to -35 dBm

Resolution: 1 dBm

Accuracy: ±1 dBm

2nd Order Receiver Filters: Centered at 520 Hz and 2240 Hz

3rd Order Receiver Filter: Centered at 1900 Hz

2nd Order Products

Range: To 50 dB below received level

Accuracy: ±1 dB

3rd Order Products

Range: To 60 dB below received level

Accuracy: ±1 dB

PEAK-TO-AVERAGE RATIO

Transmitter

Level

Range: 0 dBm to -40 dBm

Resolution: 1 dBm

Receiver

Level

Range: +10 dBm to -30 dBm

Resolution: 1 dBm

PAR

Range: 0 PAR Units to 120 PAR Units

Resolution: 1 PAR Unit

Accuracy: ±2 PAR Units

GENERAL

Transmitter

Longitudinal Balance: 60 dB to 6 kHz, decreasing 6 dB per octave above 6 kHz

Impedances: 600 and 900

Return Loss: Greater than 40 dB

SF Skip: 2450 Hz to 2750 Hz

Receiver

Longitudinal Balance: 60 dB to 6 kHz, decreasing 6 dB per octave above 6 kHz

Terminating Impedances: 600 and 900

Bridging Impedance: Greater than 50 k

Bridging Loss: Less than 0.3 dB at 1 kHz

Return Loss: Greater than 4 dB

High Frequency Protection: Greater than 60 dB attenuation above 500 kHz

Monitor Amplifier Output

Output Range: 0 volts to 7 volts

Bandwidth (3 dB points). 200 Hz to 6200 Hz

Frequency Response: ± 1 dB from 300 Hz to 3900 Hz

Output Impedance : 5

Output Connector: 347 jack

Auxiliary Outputs-BNC

Carrier: Square wave output of received frequency VCO. Nominal 10 k source impedance.

PM: AC signal proportional to band limited phase jitter. Nominal 10 k source impedance.

MISCELLANEOUS

Test Jacks: 310 jack multiplied with 5 way banding posts for both transmitter and receiver

DC Blocking (with no holding). 200 volts

Power Requirements: 105 to 129 volts AC, 60 Hz

Operating Temperature: 32 F to 122 F (0 C to 50 C)

Storage Temperature: -40 F to 167 F (-40 C to 75 C)

Warm-up Time for stated accuracy : 5 minutes

Dimensions: 13 x 15 x 22 inches

Weight: 18 pounds

Cost: $1100

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